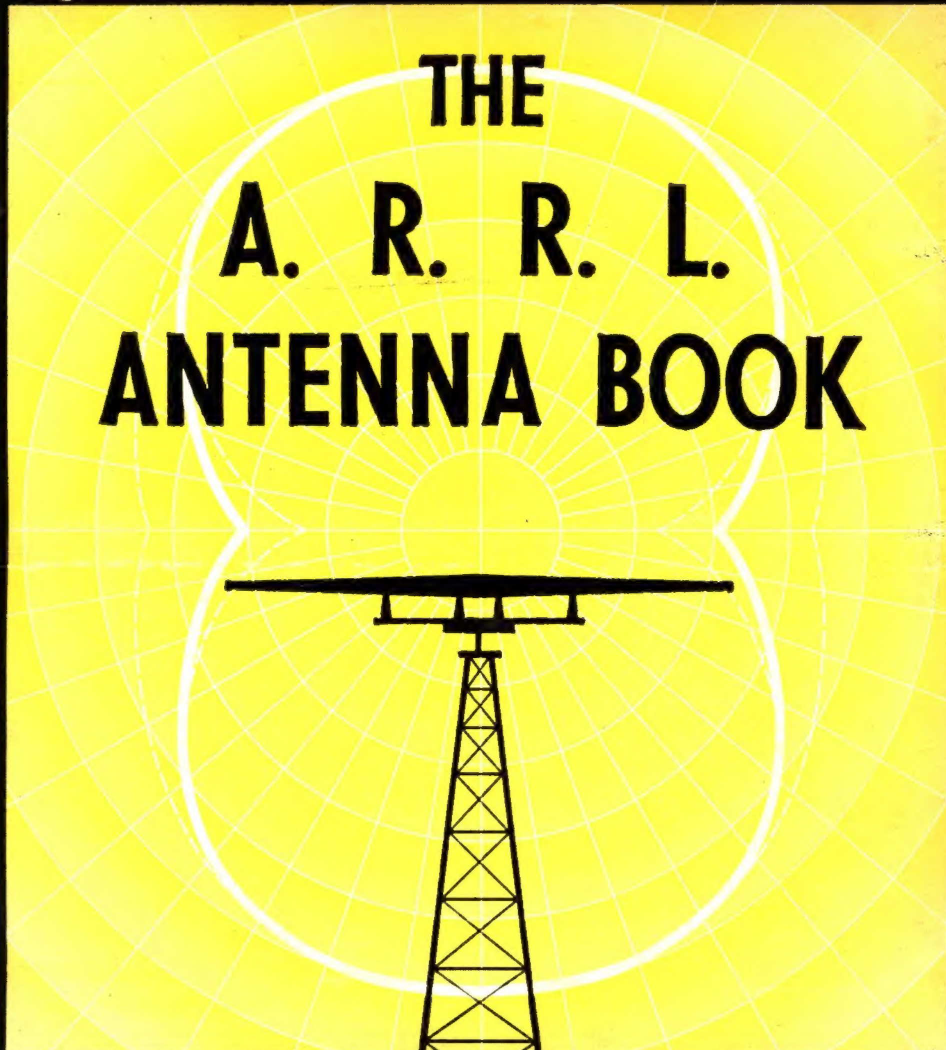
THE A. R. R. L. ANTENNA BOOK



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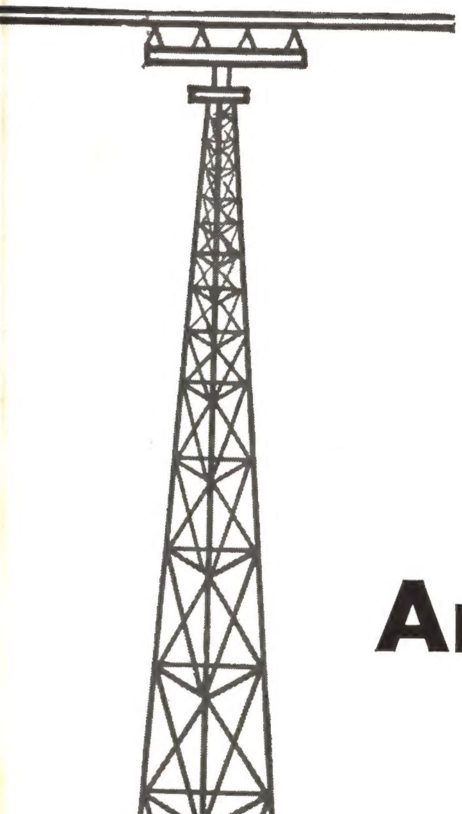

The A.R.R.L. Antenna Book

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This is a typical "antenna-raising party" at an ARRL Field Day, during which portable amateur stations go into the field to operate from temporary locations. Here the crew at K6CLZ/6 secure the guy wires of a 2-element rotary beam for 20 meters, while in the background can be seen supports for an 80-meter long wire. The low power generally used by the portable stations in Field Day makes a good site and efficient antennas of great importance.

CONTENTS

Foreword...

Radio amateurs are people with diversified interests, ranging from low frequencies to u.h.f., through radiotelephony, code operation, and teletype to television, from the sociability of "rag chewing" and the excitement of contests to handling traffic for the public benefit. In all these activities there is one common element—the antenna. It is fair to say that the ultimate success of the station is in most cases determined more by the antenna than by any other single category of equipment.

It is the purpose of this volume to assemble such of the available information on antennas as may be useful to amateurs. The subject has many ramifications, and consideration of the antenna alone is not enough: To be able to choose an antenna intelligently one must know something about how radio waves travel, because the energy radiated must be properly directed if it is to do the most good; one must also know something about transmission lines, because power can be wasted between the transmitter and antenna if the two are not properly connected. Wave propagation and transmission lines therefore require detailed treatment along with the antenna itself. There is no one "best" antenna system for all purposes, but with the help of the information in this book any amateur should be able to take the three steps that lead to a successful installation—first, to determine the things the antenna must do if it is to provide the best communication between two points; second, to choose the type of antenna that best meets the requirements; and third, to select a suitable method of transferring power from the transmitter to the antenna.

The book has two principal divisions. Chapters One through Five deal with the principles of antennas and transmission lines, wave propagation and its relationship to antenna design, and the performance characteristics of directive antenna systems. These five chapters enable the reader to design a system of his own to fit his particular needs. Beginning with Chapter Six, there is a series of chapters in which complete data are given on specific designs for the various amateur bands. The amateur who has not studied the first section, or who wishes to avoid the necessity for making his own calculations, will find in these chapters the information necessary for putting up the system that appeals to him. The remaining chapters deal with the highly-important mechanical features of construction and related subjects such as determining geographical directions.

It is sincerely hoped that you will find the work helpful. If there are some things left untreated that you wish to know more about, some things you do not understand, we shall be grateful if you, the user of the book, will give us your suggestions on how a future edition can be made even more valuable to you.

Newington, Connecticut

John Huntoon,
General Manager, ARRL

Wave Propagation

Because radio communication is carried on by means of electromagnetic waves traveling through the earth's atmosphere, it is desirable to know something about the characteristics of waves and the way in which their behavior is influenced by the conditions they meet in their trip from the radio transmitter to the receiver. While detailed knowledge of wave propagation is not at all essential to the amateur who wants to put up an effective antenna, a few facts must be understood before the principles of antenna design can be intelligently applied. An antenna may—and usually does—radiate the power applied to it with a high degree of efficiency, but if that power does not travel to the desired receiving point but goes somewhere else instead, the antenna is a failure.

Our purpose in this chapter, therefore, is to discuss those features of wave propagation that have some bearing on the design of an antenna system. In doing this we do not mean to infer that there is nothing more that the wide-awake amateur will want or need to know about the subject. Effective radio communication results from a combination of equipment, antenna system, *and* operating skill, which must include that ability to anticipate "conditions" that in turn is based on an understanding of the vagaries of wave propagation. But the latter, fascinating though they are, are somewhat outside the scope of this book when they do not directly affect antenna design.

WAVE CHARACTERISTICS

The reader who has some knowledge of electricity has already been introduced to the idea of electric and magnetic fields. A radio wave is a special combination of both types of field, with the energy divided equally between the two. If the waves could originate at a point source in "free" space—empty space such as occurs, for all practical purposes, in the interplanetary and interstellar stretches of the universe—they would spread out in ever-growing spheres with the source as the center. The speed at which the spheres expand would be the same as the speed of light, since light also is an electromagnetic wave. In empty space this speed is 300,000,000 meters per second, or approximately 186,000 miles in one second. The path of a ray from the source to any point on the spherical surface always is a straight line—a radius of the sphere.

It is obvious that in a remarkably short time a sphere growing outward from the center would be very large indeed. An observer on such a spherical surface would conclude, if he could "see" the wave in his vicinity, that it does not appear to be spherical at all but instead seems like a flat surface—just as the earth seems to human beings to be flat rather than spherical. A wave that is far enough from the course to appear flat is called a plane wave. The radio waves with which we deal in communication always meet this condition, at least after they have traveled a short distance from the transmitting antenna.

A typical representation of the lines of electric and magnetic force in a plane wave is given in Fig. 1-1. The nature of wave propagation is such

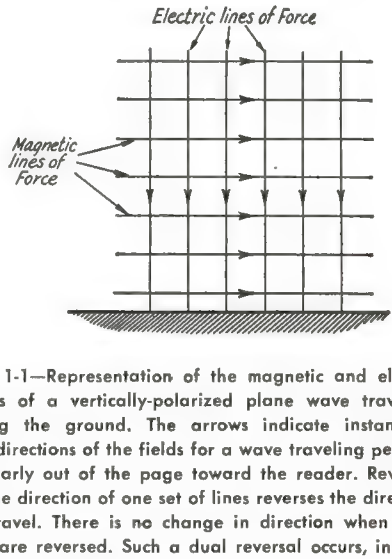


Fig. 1-1—Representation of the magnetic and electric fields of a vertically-polarized plane wave traveling along the ground. The arrows indicate instantaneous directions of the fields for a wave traveling perpendicularly out of the page toward the reader. Reversal of the direction of one set of lines reverses the direction of travel. There is no change in direction when both sets are reversed. Such a dual reversal occurs, in fact, once each half cycle.

that the electric and magnetic lines always are mutually perpendicular, as indicated in the drawing. The plane containing the set of crossed lines represents the wave front. The direction of wave travel always is perpendicular to the wave front, but whether the direction is "forward' or "backward" is determined by the relative directions of the electric and magnetic forces.

If the wave is traveling through anything other than empty space its speed is not 300,000,000 meters per second but is something less. Just how much less depends on the substance or medium through which the wave is traveling. If the medium is air instead of empty space, the reduction in speed is so small that it can be ignored in most calculations. In solid insulating materials the speed is in general much slower;

for example, in distilled water (which is a good insulator) the waves travel only one-ninth as fast as they do in space. In good conductors such as metals the speed is so low that the opposing fields set up by currents induced in the conductor by the wave itself occupy practically the same space as the original wave and thus almost cancel it out. This is the reason for the skin effect in conductors at high frequencies and also the reason why thin metal enclosures form good shields for electrical circuits at radio frequencies.

Phase and Wavelength

Because the speed at which radio waves travel is so great, we are likely to fall into the habit of ignoring the time that elapses between the instant at which a wave leaves the transmitting antenna and the instant at which it arrives at the receiving antenna. It is true that it takes only one-seventh of a second for a wave to travel around the world, and from a communication standpoint that is hardly worth worrying about. But there is another consideration that makes this factor of *time* extremely important.

The wave is brought into existence because an alternating current flowing in a conductor (which is usually an antenna) sets up the necessary electric and magnetic fields. The alternating currents used in radio work may have frequencies anywhere from a few tens of thousands to several billion cycles per second. Suppose that the frequency is 30 megacycles per second —that is, 30,000,000 cycles per second. One of those cycles will be completed in 1/30,000,000 second, and since the wave is traveling at a speed of 300,000,000 meters per second it will have moved only 10 meters during the time that the current is going through one complete cycle. To put it another way, the electromagnetic field 10 meters away from the antenna is caused by the current that *was* flowing in the antenna *one cycle earlier* in time; the field 20 meters away is caused by the current that *was* flowing *two* cycles earlier, and so on.

Now if each cycle of current is simply a repetition of the one that preceded it, the currents at corresponding instants in each cycle will be identical, and the fields caused by those identical currents also will be identical. As the fields move outward they become more thinly spread over larger and larger surfaces, so their amplitudes decrease with distance from the antenna. But they do not lose their identity with respect to the instant of the cycle at which they were generated. That is, the phase of the outwardly-moving surface remains constant. It follows, then, that at intervals of 10 meters (in the example above) measured outward from the antenna the phase of the waves at any given instant is identical.

In this fact we have the means for defining rather precisely both "wave front" and "wavelength." The wave front is simply a surface in every part of which the wave is in the same phase. The wavelength is simply the distance between two wave fronts having identical phase at any given instant. In the example, the wavelength is 10 meters because the distance between two wave fronts having the same phase is, as we found, 10 meters. This distance, incidentally, always must be measured perpendicular to the wave fronts; in other words, along the same line that represents the direction in which the wave is traveling. Measurements made along any other line between the two wave fronts would lead to the erroneous conclusion that the wave is longer than it really is. Expressed in a formula, the length of a wave is

$$\lambda = \frac{v}{f}$$

where λ = wavelength
v = velocity of wave
f = frequency of current causing the wave

The wavelength will be in the same length units as the velocity when the frequency is expressed in the same time units as the velocity. For waves traveling in free space (and nearly enough for waves traveling through air) the wavelength is

$$\lambda\,(\text{meters}) = \frac{300}{f\,(\text{Mc.})}$$

We shall continually be encountering this idea of phase in succeeding chapters in this book, because it is fundamental to the operation of antenna systems. It is essential, therefore, to have a clear understanding of what it means if antenna behavior is to be appreciated. Basically, "phase" means "time," but when something goes through periodic variations with time in the way that an alternating current does, corresponding instants in succeeding cycles are said to have the same phase even though the actual time difference is one cycle. In using the word phase in this fashion we are inherently using the cycle as the unit of time measurement, just as we use one 24-hour day as a unit of time measurement. Four o'clock this afternoon corresponds to four o'clock yesterday afternoon in much the same way that an instant in an a.c. cycle corresponds to the identical instant in the preceding cycle.

In Fig. 1-2 points *A*, *B* and *C* are all in the same phase because they are corresponding instants in each cycle. This is a conventional drawing of a sine-wave alternating current with time progressing to the right. It also represents an instantaneous "snapshot" of the distribution of intensity of the traveling fields if distance is substituted for time in the horizontal axis. In that case the distance between *A* and *B* or between *B* and *C* represents one wavelength. This shows that the field-intensity distribution follows the sine curve, both as to amplitude and polarity, to correspond exactly to the time variations in the current that produced the fields. It must be remembered that this is an *instantaneous* picture; the actual wave travels along just as a wave in water does. To an observer at any fixed point along the wave path, the field intensity goes through *time* variations corresponding to the

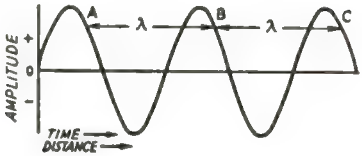


Fig. 1-2—The instantaneous amplitude of both fields (electric and magnetic) varies sinusoidally with time as shown in this graph. Since the fields travel at constant velocity, the graph also represents the instantaneous distribution of field intensity along the wave path. The distance between two points of equal phase, such as A-B and B-C, is the length of the wave.

time variations of the current that initiated the wave.

Field Intensity

The strength of a wave is measured in terms of the voltage between two points lying on an electric line of force in the plane of the wave front. The unit of length is the meter, and since the voltage in a wave usually is quite low, the measurement is made in microvolts per meter. The voltage so measured goes through time variations just like those of the original current that caused the wave, and so is measured like any other a.c. voltage—that is, in terms of the effective value or, sometimes, the peak value.

There are few, if any, occasions in amateur work where a measurement of actual field strength is necessary. This is fortunate, because the equipment required is elaborate. It is comparatively easy, however, to make measurements of *relative* field strength, and thus determine whether an adjustment to an antenna system has resulted in an improvement or not.

Polarization

A wave such as is shown in Fig. 1-1 is said to be polarized in the direction of the electric lines of force. In this drawing the polarization is vertical because the electric lines are perpendicular to the earth. A wave "with its feet on the ground" as in Fig. 1-1 is, as a matter of fact, usually vertically polarized. This is because the ground acts as a rather good conductor, particularly at frequencies below about 10 Mc., and it is one of the laws of electromagnetic action that electric lines touching the surface of a good conductor must do so perpendicularly. Over partially conducting ground there may be a forward tilt to the wave front; this tilt in the electric lines is greater as the energy loss in the ground becomes greater.

Waves traveling in contact with the surface of the earth are of little usefulness in amateur communication because as the frequency is raised the distance over which such a "surface" wave will travel without excessive loss of energy or attenuation becomes smaller and smaller. The surface wave is of most utility at low frequencies and through the standard b.c. band. At high frequencies the wave reaching the receiving antenna ordinarily has not had much contact with the earth and its polarization is not necessarily vertical. If the electric lines of force are horizontal, the wave is said to be horizontally polarized. However, the polarization can be anything between horizontal and vertical. In many cases, the polarization is not fixed but continually rotates. When this occurs the wave is said to be elliptically polarized.

Attenuation

In free space the field intensity of the wave decreases directly with the distance from the source. That is, if the field strength one mile from the source has a value, let us say, of 100 microvolts per meter, the field strength at two miles will be 50 microvolts per meter, at 100 miles will be 1 microvolt per meter, and so on. This decrease in field strength is caused by the fact that the energy in the wave has to spread out over larger and larger spheres as the distance from the source is increased.

In actual communication by radio the attenuation of the wave may be much greater than this "inverse-distance" law would indicate. For one thing, the wave is not traveling in empty space. For another, the receiving antenna seldom is situated so that there is a clear "line of sight" between it and the transmitting antenna. Since the earth is spherical and the waves do not penetrate its surface to any considerable extent, communication has to be by some means that will bend the waves around the curvature of the earth. These means exist, but they usually involve additional energy losses that increase the attenuation of the wave with distance.

Reflection, Refraction and Diffraction

It has been mentioned that radio waves and light waves are the same type of wave; the only difference is in the scale of wavelength. We are all familiar with the reflection of light; radio waves are reflected in much the same way. Frequently, however, the reflecting surfaces are small (in terms of wavelength) compared with the surfaces from which we see light waves reflected. An object the size of an automobile, for instance, will not reflect much of the energy in an 80-meter wave. On the other hand, it may be a very good reflector of waves only a meter or two in length. The thickness of the object is of some importance because the waves penetrate it to an extent depending on its characteristics. In a material of given conductivity, for example longer waves will penetrate farther than shorter ones and so require a greater thickness for good reflection. Thin metal is a good reflector even at quite long wavelengths, but in poorer conductors such as the earth—which certainly meets the requirement of having a surface that is large compared with any radio wavelength—the longer wavelengths may penetrate quite a few feet.

Reflection may also take place from any surface that represents a change in the dielectric

constant of the medium in which the wave is moving. A familiar example in optics is the reflection of light from the surface of a pane of glass that is itself quite transparent to light waves. When viewed from certain angles, it is practically impossible to see through the pane of glass because of the reflected light.

Another phenomenon that has a rather familiar counterpart in optics is refraction, or the bending that takes place when the wave enters (at an angle) a medium having a different dielectric constant than the medium it has just left. This bending is caused by the fact that the wave travels at a different speed when the dielectric constant is changed. The part of the wave that enters the new medium first is either slowed down or speeded up, depending on the relative dielectric constants, and so tends to get ahead of or fall behind the sections of the wave that enter later. The effect is to change the direction in which the wave is moving. The classic example in optics is the apparent sharp bend in a stick held partly in and partly out of a body of water at an angle.

Most of the optical examples of refraction are based on two homogeneous substances having a very definite common boundary, as between air and glass. In that case the rays travel in straight lines *inside* either medium and the bending takes place at the common surface. In radio transmission it is frequently the case that the boundary between the two areas of differing dielectric constant is not at all sharp; the dielectric constant simply changes gradually over quite a distance along the wave path. This causes the wave bending also to be gradual, and the wave path becomes curved.

A somewhat less familiar optical phenomenon that has its radio counterpart is diffraction. To the eye, the shadows cast by a pin-point source of light appear to be quite sharp. However, close examination shows that light bends around the edge of an object to some extent, depending on the thickness of the edge. This effect becomes greater as the wavelength is increased, and can be of some importance at radio frequencies. For example, with waves traveling in a straight line one would expect that no signal could be heard behind a hill, but the bending caused by diffraction does produce a signal in the "shadow area." At high radio frequencies the diffracted signal is weak compared with the direct ray, and frequently is masked by stronger signals reaching the same spot by other means such as reflection or refraction in the atmosphere.

Both reflection and refraction can take place in various parts of the atmosphere, and the mechanisms by which they occur are likewise varied. The result is that radio waves frequently are "scattered," just as light is scattered in the atmosphere. Such scattering accounts for the reception of signals under conditions when they would not be expected from the simplified pictures of wave travel now to be discussed.

The Ground Wave

Waves travel close to the earth in several different ways, some of which involve relatively little contact with the ground itself. The selection of proper nomenclature therefore becomes somewhat confusing, but more or less by common consent the term ground wave is applied to waves that stay close to the earth and do not reach the receiving point by reflection or refraction from the much higher region of the atmosphere known as the "ionosphere." The ground wave therefore can be a wave traveling in actual contact with the ground, such as the wave pictured in Fig. 1-1, or it can be a wave that goes directly from the transmitting antenna to the receiving antenna when the two antennas are high enough so that they can "see" each other. It can also be a wave that is refracted or reflected in the atmosphere near the earth (the troposphere).

THE SURFACE WAVE

A wave that travels in contact with the earth's surface is called a surface wave. It is the type of wave that provides reception up to distances of 100 miles or more in the standard broadcast band during the daytime. The attenuation of this type of wave is rather high, so the intensity dies off rapidly with distance from the transmitter. The attenuation also increases rapidly with frequency, with the result that the surface wave is of little value in amateur communication with the possible exception of distances up to perhaps 50 miles on the 3.5-Mc. band. As explained earlier, the surface wave must be essentially vertically polarized. The transmitting and receiving antennas therefore must generate and receive vertically-polarized waves, if the surface wave is to be utilized to advantage. In general terms, this means that both antennas must be vertical.

THE SPACE WAVE

The conditions that exist when the transmitting antenna and receiving antenna are within line of sight of each other are shown in Fig. 1-3. One ray travels directly from the transmitter to the receiver and consequently is attenuated in about the same way as a wave in free space. However, the wave from the transmitting antenna also strikes the ground between the two antennas, and the ray that does so at the proper angle to reach the receiving antenna (the angle of incidence being equal to the angle of reflec-

tion, as in optics) combines with the direct ray to produce the actual signal at the receiving antenna.

Fig. 1-3—The ray traveling directly from the transmitting antenna to the receiving antenna combines with a ray reflected from the ground to form the space wave.

In most practical cases where the communication is between two stations on the ground (as contrasted with communication between ground and an airplane, or between two airplanes) the angle at which the ground-reflected ray strikes the earth and is reflected will be very small. That is, the ray strikes the earth at almost grazing incidence. Now it happens that such a reflection reverses the phase of the wave, so if the distance traveled by the direct wave and the distance traveled by the ground-reflected ray were exactly the same, the two rays would arrive out of phase and would cancel each other. Actually, the ground-reflected ray has to travel a little farther, and so the phase difference between the two rays depends on the difference in path length as measured in terms of wavelength. If the difference in the length of the two paths is 3 meters, for example, the phase difference from this cause will be only 3 degrees if the wave is 360 meters long. This is only a negligible shift in phase from the 180 degrees caused by the reflection, and so the signal strength would be small. On the other hand, if the wavelength is 6 meters the phase shift caused by the same difference in path length would be 180 degrees—enough to overcome completely the 180-degree reversal caused by the reflection, so the two rays would add at the receiving antenna. In short, the space wave is a negligible factor in communication at *low* frequencies, because the difference in the distance traveled by the two rays always is very small, when measured in terms of wavelength. The space wave therefore is canceled out at such frequencies. But as the frequency is raised (wavelength shortened) the space wave becomes increasingly important. It is the dominating factor in ground-wave communication at v.h.f. and u.h.f.

The space-wave picture presented is a simplified one and, as usual, there are practical complications that modify it. There is some loss of energy when the ray strikes the ground, so the reflected ray does not arrive at the receiving antenna with the same intensity as the direct ray. Because of the ground loss the phase of the reflected ray is not shifted exactly 180 degrees. For both these reasons the two waves never cancel completely at the receiving antenna. Also, at frequencies in the u.h.f. region it is possible to form the wave into a beam, much like the light beam from a flashlight. Such a beam puts most of the energy into the direct ray and reduces the amount that can strike the ground, particularly when the transmitting and receiving antennas are both at high elevations. Thus the effect of the ground-reflected ray is minimized.

Strictly speaking, the description above applies only to a horizontally-polarized wave and perfectly-conducting earth. Practically, the polarization does not make much difference because the earth is neither a perfect conductor nor a perfect dielectric. The overall result is that at frequencies below, say, 20 Mc., the space wave is inconsequential. But at v.h.f. it is readily possible to transmit to the horizon by means of the space wave.

"Line-of-Sight" Propagation

From inspection of Fig. 1-3 it appears that use of the space wave for communication between two points depends on having a line of sight between the two locations. This is not quite literally true. The structure of the atmosphere near the earth is such that under "normal" conditions (a theoretical normal, rather than an actual one; in many parts of the world, at least, the "normal" is an average which is statistically useful but seldom represents the actual condition of the atmosphere) the waves are bent into a curved path that keeps them nearer to the earth than true straight-line travel would. This effect can be aproximated by assuming that the waves travel in straight lines but that the earth's radius is increased in dimension by one-third. On this assumption, the distance from the transmitting antenna to the horizon is given by the following formula:

$$D\ (\text{miles}) = 1.41\ \sqrt{H\ (\text{feet})}$$

where H is the height of the transmitting antenna, as shown in Fig. 1-4. The formula assumes

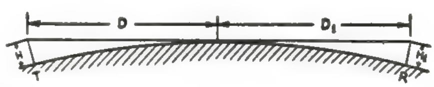


Fig. 1-4—The distance, *D*, to the horizon from an antenna of height *H* is given by the formula in the text. The maximum line-of-sight distance between two elevated antennas is equal to the sum of their distances to the horizon, as indicated in the drawing.

that the earth is perfectly smooth out to the horizon; of course, any obstructions that rise along any given path must be taken into consideration. The point at the horizon is assumed to be on the ground. If the receiving antenna also is elevated, the maximum line-of-sight distance between the two antennas is equal to $D + D_1$; that is, the sum of the distance to the horizon from the transmitting antenna and the distance to the horizon from the receiving antenna. The

distances are given in chart form in Fig. 1-5. Two stations on a flat plain, one having an antenna on a tower 60 feet high and the other having an antenna supported 40 feet in the air, could be separated approximately 20 miles for line-of-sight communication.

In addition to the "normal" refraction or bending, the waves also are diffracted around the curvature of the earth, so that the actual distance that can be covered does exceed the line-of-sight distance. However, under ordinary conditions the amount of diffraction at v.h.f. and u.h.f., where the space wave is of chief importance, is rather small and the signal strength drops off very rapidly in a short distance beyond the earth's "shadow."

To make maximum use of the ordinary space wave discussed here it is necessary that the antenna be as high as possible above the surrounding country. A hill that juts above the adjacent

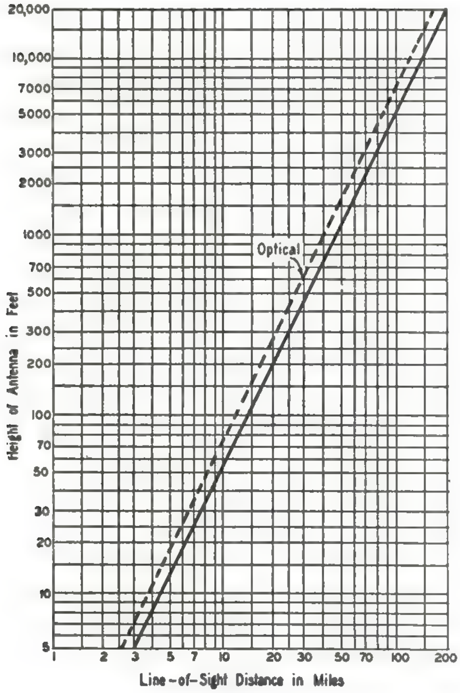


Fig. 1-5—Distance to the horizon from an antenna of given height. The solid curve includes the effect of atmospheric refraction. The optical line-of-sight distance is given by the broken curve.

terrain is usually an excellent location. However, the peak of a hill is not necessarily the best spot, particularly if it is of the nature of a plateau. Arriving waves may have to be diffracted over the brow of the hill to reach the antenna unless the latter is placed on a high pole or tower; in other words, the brow of the hill may shield the antenna from waves arriving from a desired direction. Also, it is advantageous to have the ground drop off fairly sharply in front of the antenna, as this frequently prevents the ground-reflected ray from approaching at such a flat angle as it would over level ground. Generally speaking, a location just below the peak of a hill is the optimum one for transmitting and receiving in a desired direction, as indicated in Fig. 1-6.

Since the space wave goes essentially in a straight line from the transmitter to the receiver, the antenna used for radiating it should concentrate the energy toward the horizon. That is, the antenna should be a "low-angle" radiator, because energy radiated at angles *above* the horizon obviously will pass over the receiving antenna. Similarly, the receiving antenna should be most responsive to waves that arrive horizontally.

In general, the polarization of a space wave remains constant during its travels. Therefore, the receiving antenna should be designed to give maximum response to the polarization set up at the transmitting antenna. For v.h.f. work both horizontal and vertical polarization are used, the former being more generally preferred. The principal reason for this preference is that the chief source of radio noise at v.h.f.—that generated by the spark in the ignition systems of automobiles—is predominantly vertically polarized. Thus horizontally-polarized antennas tend to discriminate against such noise and thereby improve the signal-to-noise ratio.

PROPAGATION IN THE TROPOSPHERE

Weather conditions in the atmosphere at heights from a few thousand feet to a mile or two at times are responsible for bending waves downward. This tropospheric refraction makes communication possible over far greater distances than can be covered with the ordinary space wave. The amount of the bending increases with frequency, so tropospheric communication improves as the frequency is raised. The bending is relatively inconsequential at frequencies below 28 Mc., but provides interesting communication possibilities at 50 Mc. and above.

Refraction in the troposphere takes place when masses of air become stratified into regions having differing dielectric constants. If the boundary between the two masses of air is sharply defined, reflection as well as refraction may take place for waves striking the boundary at grazing angles. The most common cause of tropospheric refraction is the temperature inversion. Normally, the temperature of the lower atmosphere decreases at a constant rate of approximately 3 degrees F. per 1000 feet of height. When this rate is decreased for any reason, a temperature inversion is said to exist and greater-than-normal wave bending takes place. Some of the types of temperature inversion are the *dynamic* inversion, resulting when a warm air mass overruns a colder mass; the *subsidence* inversion, caused by the sinking of an air mass heated by compression; the *nocturnal* inversion, brought about by the rapid cooling of surface air after

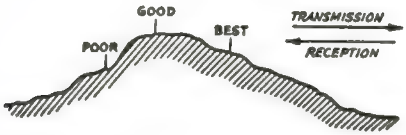


Fig. 1-6—Propagation conditions are generally best when the antenna is located slightly below the top of a hill on the side that faces the distant station. Communication is poor when there is a sharp rise immediately in front of the antenna in the direction of communication.

sunset; and the *cloud-layer* inversion, caused by the heating of air above a cloud layer by reflection of the sun's rays from the upper-surface of the clouds. Sharp transitions in the water-vapor content of the atmosphere may also bring about refraction and reflection of v.h.f. waves.

Because the atmospheric conditions that produce tropospheric refraction are seldom stable over any considerable period of time, the strength of the received signal usually varies or "fades" over a wide range. Hourly and seasonal variations are observed. Best conditions often occur in the evening and just before sunrise, and conditions are generally poorest at midday when the atmosphere is relatively stable. Tropospheric refraction is generally greatest in the early summer and early fall. It is also more pronounced along the seacoasts.

The tropospheric wave maintains essentially the same polarization throughout its travel, so the transmitting and receiving antennas should have the same type of polarization. Since waves that enter the refracting region at anything other than practically grazing incidence are not bent enough to be useful for communication, the transmitting antenna should be designed for maximum radiation horizontally. The receiving antenna likewise should be a low-angle affair if the received signal is to be most efficiently utilized.

Atmospheric Ducts

In some parts of the world, particularly in the tropics and over large bodies of water, temperature inversions are present practically continuously at heights of the order of a few hundred feet or less. The boundary of the inversion is usually well enough defined so that waves traveling horizontally are "trapped" by the refracting layer of air and continually bent back toward the earth. The air layer and the earth form the upper and lower walls of a "duct" in which waves are guided in much the same fashion as in a metallic wave guide. The waves therefore follow the curvature of the earth for distances (sometimes hundreds of miles) far beyond the optical horizon of the transmitter.

Because the height of an atmospheric duct is relatively small, only waves smaller than a certain limit wil be trapped. If the refracting layer is only a few feet above the surface the lowest usable frequency may be as high as a few thousand megacyles, so ultrahigh or superhigh frequencies must be used. Under some conditions, however, the height and dielectric characteristics of the layer may be such that waves in the medium v.h.f. region will be transmitted. The line of distinction, if any, between ducting and ordinary tropospheric propagation is hard to draw in such a case.

A feature of duct transmission is that the antennas, both transmitting and receiving, must be *inside* the duct if communication is to be established. If the duct extends only a few feet above the earth and the transmitting antenna is on a tower or promontory *above* the duct, no signals will be heard at the receiving point. Likewise, a receiving antenna situated above the duct will not pick up energy trapped nearer the earth.

Atmospheric ducts also are formed between two layers of air having suitable characteristics. If the lower layer refracts the waves *upward* while the upper layer refracts them *downward*, waves will be trapped between the two layers and again can travel for great distances. In such a case antennas either below or above the duct will be ineffective. Ducts of this type have been observed from airplanes, where good signals will be received with the plane at the optimum height but the signal strength drops off rapidly at either higher or lower altitudes.

Much remains to be learned about the extent of duct transmission at amateur frequencies. There appears to be no significant difference in the signal strength with either horizontal or vertical polarization. Communication via ducts may turn out to be the most important mode at u.h.f.

Other Modes

Although the transmission modes just discussed are the most common ones at v.h.f., they are by no means the only methods by which communication can be carried on at such frequencies. New modes are continually being discovered as better antennas and equipment are developed. However, it is invariably true that best results in all types of v.h.f. propagation are obtained when the transmitting antenna system concentrates as much as possible of its energy in the very low vertical angles, and when the receiving antenna likewise is most responsive to signals arriving horizontally.

The Sky Wave

At frequencies below 30 Mc. practically all amateur communication—except for "local" work over distances of a few miles—is carried on by means of the sky wave. This is a wave that, on leaving the transmitting antenna, would travel on out into empty space if it were not for the fact that under certain conditions it can be sufficiently reflected or refracted, high up in the earth's atmosphere, to reach the earth again at distances varying from zero to about 2500 miles

from the transmitter. By successive reflections at the earth's surface and in the upper atmosphere, communication can be established over the maximum possible terrestrial distances.

THE IONOSPHERE

The region in which the waves are bent back to earth is called the ionosphere. This is a section of the upper atmosphere in which the air pressure is so low that "free" electrons and ions can move for a long time without getting close enough to each other to be attracted together and thus recombine into a neutral atom. A wave entering a region in which there are many free electrons will be affected in much the same way as one entering a region of differing dielectric constant; that is, its direction of travel will be shifted. The mechanism is complicated, but in a broad sense is the result of the interactions that take place when the free electron is set in motion by electromagnetic forces of the passing wave. In the ionosphere the moving wave tends to be bent back toward the earth.

Ultraviolet light from the sun is the primary cause of ionization in the upper atmosphere. The amount of ionization does not change uniformly

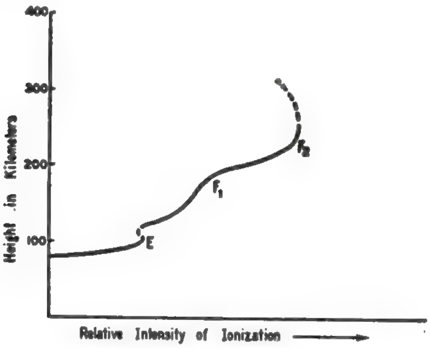


Fig. 1-7—Typical curve of distribution of ion density with height in latitude of Washington, D. C., near a sunspot minimum. The ion density and the heights at which the maxima occur for each layer will depend on the latitude and the period in the sunspot cycle. The D region is not shown because it is not ordinarily measurable by the methods used in ionosphere sounding; its presence is known principally because of the observed absorption of lower-frequency waves during the daytime hours.

with height above the earth, as might be expected at first thought. Instead, it is found that there are relatively dense layers of ionization, quite thick vertically, at rather well-defined heights. Nor is the ionization uniform within the layer itself; it is highest at the center of the layer and tapers off gradually both above and below. Fig. 1-7 is a representative plot of the intensity of ionization with height above the earth. Both the height and ionization intensity of any given region vary with the time of day, the season of the year, and the 11-year sunspot cycle. This is because the amount of ultraviolet radiation received from the sun at any given spot depends on these factors.

Layer Characteristics

The ionized layers or regions are designated by letters. The lowest one known, at a height of about 30 to 55 miles, is called the *D* region. Because it is in a relatively dense part of the atmosphere the atoms broken up into ions by sunlight quickly recombine, so the amount of ionization depends directly on the amount of sunlight. Thus *D*-region ionization is maximum at local noon and disappears at sundown. When electrons in the *D* region are set in motion by a passing wave the collisions between particles are so frequent, because of the rather high air density, that a substantial proportion of the wave energy may be used up as heat. The probability of collisions depends on the distance an electron can travel under the influence of the wave. This distance depends on the frequency of the wave, because during a long cycle (low frequency) the electron has time to move farther, before the direction of the field reverses and sends it back again, than it does in a short cycle (high frequency). If the frequency is low enough the collisions between particles will be so frequent that practically all energy in the wave will be absorbed in the *D* region. This usually happens at frequencies in the 3.5-4.0 Mc. amateur band at the time of maximum *D*-region ionization, particularly for waves that enter the layer at the lower vertical angles and thus have to travel a relatively long distance through it. At times of sunspot maxima even waves entering the layer directly upward will be almost wholly absorbed, in this frequency band, around the middle of the day. The absorption is less in the 7-Mc. band and is quite small at 14 Mc. and higher frequencies. The *D* region is relatively ineffective in bending high frequency waves back to earth, and so plays no significant part in amateur long-distance communication except as an absorber of energy. It is the principal reason why daytime communication on the lower frequencies (3.5 and 7 Mc.) is confined to short distances.

The lowest layer that affords long-distance communication has a mean height of about 65 miles and is called the *E* layer. It is a region of fairly high atmospheric density and consequently the ionization varies with the height of the sun. The ionizaton drops rapidly after sundown, when the ions and electrons recombine in the absence of sunlight, and reaches a minimum at about midnight. It again increases rapidly at sunrise and reaches a maximum at about noon local time. As in the case of the *D* region, the *E* layer absorbs energy from low-frequency waves during the time of maximum ionization.

The second important "communication" layer is the F_2 layer. This is the most intensely ionized layer, and its height—of the order of 150 to 250 miles—varies with the time of day, the season, and the sunspot cycle. At these heights the at-

mosphere is very thin, and so the ions and electrons are slow to recombine. Because of this, the ionization is not so responsive to the height of the sun; it reaches a maximum shortly after noon, local time, but "tails off" gradually thereafter. It continues at a rather high, but gradually decreasing level throughout the night, reaching a minimum just before sunrise. At sunrise it increases rapidly and attains the daytime level in the course of an hour or two.

During the day the F_2 layer sometimes splits into two layers, the lower (and weaker) one occurring at a height of 120 miles or so and being designated the F_1 layer is, in general, of little importance in communication, except for introducing additional energy absorption of waves traveling through it. It disappears at night. After sundown, too, the height of the F_2 layer decreases, the maximum ionization occurring in the neighborhood of 175 miles. (The nighttime F_2 layer was formerly called the F layer, but present usage is to drop the separate designation for the night layer.)

Refraction in the Ionosphere

The amount by which the path of a wave is bent in an ionized layer depends on the intensity of ionization and the wavelength. The greater the ionization, the more the bending at any given frequency. Or, to put it another way, for a given degree of ionization the bending will be greater as the frequency of the wave is lowered—in other words, as its wavelength is increased.

Two extremes thus become possible. If the ionization is intense enough and the frequency is low enough, a wave entering the ionized region *perpendicularly* will be turned back to earth. But as the frequency is raised or the ionization is decreased, a condition will eventually be reached where the bending will not be sufficient to return the wave to earth, even though the wave leaves the transmitting antenna at the lowest possible angle and thus requires the least bending in the ionosphere. A typical "in-between" condition is illustrated in Fig. 1-8, a simplified illustration of the paths taken by high-frequency waves and considering only the effect of a single layer.

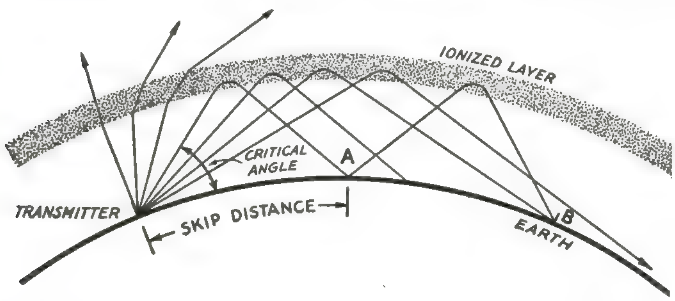


Fig. 1-8—Behavior of waves on encountering the ionosphere. Waves entering the ionized region at angles higher than the critical angle are not bent enough to be returned to earth. Waves entering the critical angle reach the earth at increasingly greater distances as the angle approaches the horizontal.

Fig. 1-8 shows a condition that is frequently typical of the way waves are bent in a single layer. (When several layers are involved, the paths are naturally more complex, since the layers have differing characteristics.) In this case the layer is capable of refracting waves that enter it at low angles. However, as the angle at which the ray strikes the layer is increased, a critical angle is reached at which the ray just manages to be bent back to earth. Rays entering at still greater angles are not bent enough and pass through the layer into empty space. Since such rays are useless for communication, it is obvious that energy radiated at angles above the critical angle is wasted.

Note also that the point at which a ray reaches the earth on its return journey from the ionosphere depends on the angle at which it left the transmitting antenna. The larger the angle with the surface of the earth the shorter the distance from the transmitter to the point at which the returning ray arrives.

Skip Distance

When the critical angle is less than 90 degrees the highest-angle wave that can be bent back to earth will return at an appreciable distance from the transmitter. For some distance, then, depending on the critical angle, there is a region about the transmitter where the sky-wave signal will not be heard. This "silent" region, extending from the limit of the useful ground wave to the distant point where the sky-wave signal can first be heard, is called the skip zone, because all signals skip over it. The skip zone is indicated by the skip distance in Fig. 1-8.

The skip distance—the distance from the transmitter to the point where the sky-wave signal is first heard—depends on the critical angle and the layer height. The lower the critical angle the farther the skip distance extends. Since higher frequencies are, in general, bent less than lower frequencies, the skip distance is greater the higher the frequency. For a given critical angle, it is also greater the greater the height of the layer in which the bending takes place. Thus for the same critical angle, the skip distance with F_2-layer bending will be greater than for waves returned to earth from the E layer, because the F_2 layer is higher.

When waves at any and all angles are returned to earth from the layer there is, of course, no skip zone. In such instances the sky wave frequently is stronger than the ground wave, at amateur frequencies, even as close as a few miles from the transmitter location. This is because the wave is attenuated less in its travel up to the

layer and back again than it is in going a few miles over the ground, surprising as it may seem.

Single- and Multihop Propagation

Fig. 1-8 also shows two of the modes by which the signal can reach a distant receiving point. In one case the wave is bent in the layer at a point about midway between the transmitter and the receiving point, B. The wave thus makes the trip in one "hop." However, that is not the only possibility. A ray that is reflected midway between the transmitter and point A (which in turn is midway between the transmitter and B) will be reflected when it strikes the earth at A and will go up to the layer again. Here it is once more reflected, returning to earth, finally, at B. This is "two-hop" transmission. More than two hops are readily possible.

Multihop propagation over long distances tends to become more complex than the simple geometrical picture given above would indicate, partly because different ionospheric conditions usually exist at each point of reflection. Observation indicates, however, that the hops are well defined over distances up to several thousand miles, and antennas for commercial point-to-point circuits usually are designed to radiate and receive best at the vertical angles associated with the number of hops between the transmitting and receiving points. The smallest possible number of hops is best, in general, since each additional hop introduces additional attentuation because of losses at the reflection points. Even more important, however, is the effect of the ionosphere itself as described later. Ionospheric absorption may be so much less, at a given operating frequency, with say three or four hops instead of two over a given path that the received signal will be much stronger in spite of the additional reflection loss.

Virtual Height and Critical Frequencies

By using a frequency low enough so that waves entering the ionosphere at the maximum angle of 90 degrees (i.e., waves going vertically from the transmitting antenna to the ionosphere) are returned to earth, it is possible to measure the height of the ionosphere. This is done by measuring the time taken by the wave to go up and back. Knowing the time and velocity of propagation, the distance can be readily calculated. The distance so found is the *virtual height*, or the height from which a pure reflection would give the same effect as the refraction that actually takes place. The method is illustrated in Fig. 1-9. Because a certain amount of time is required for the wave to make the turn at the top of its travel, the virtual height is somewhat higher than the actual height, as the illustration shows.

If the transmitting frequency is gradually increased while height measurements of this type are being made, eventually a frequency range will be encountered where the virtual height seems to increase rapidly, until finally the wave does not come back. The highest frequency that is returned to earth is known as the critical frequency. As the frequency is further increased beyond the critical frequency, the wave must enter the ionosphere at progressively smaller angles in order for it to be bent back to earth. By using very low angles long-distance transmission via the F_2 layer is possible at frequencies up to about 2.5 times the critical frequency. Thus, the critical frequency is a measure of the reflecting ability of the ionosphere.

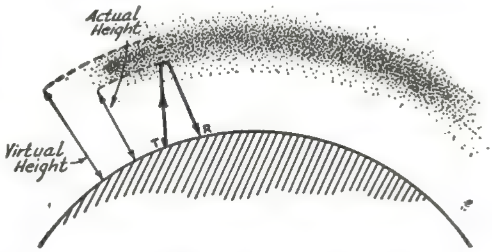


Fig. 1-9—The "virtual" height of the refracting layer is measured by sending a wave vertically to the layer and measuring the time it takes for it to come back to the receiver. The actual height is somewhat less because of the time required for the wave to "turn around" in the ionized region.

Since the refracted wave acts as though it were reflected from a mirror at the virtual height, it is customary to use the terms "reflection" and "refraction" almost interchangeably in connection with ionospheric propagation. In most cases the actual process is refraction. However, it *is* possible for true reflection to occur if the boundary of the layer is sharply defined and the wave strikes it at a small enough angle.

Virtual heights, of course, depend on the height of the ionized region. The critical frequencies vary with the intensity of ionization in the layers, being greater when the ionization increases. Since the ionization is greatest at the peak of the sunspot cycle, critical frequencies are highest in both the E and F_2 layers during that period. Conversely, they are lowest during a sunspot minimum. The E-layer critical frequency ranges from about 1 to 4 megacycles, depending on the period in the sunspot cycle and the time of day. The F_2 critical frequency varies with the time of day, the season, and the sunspot cycle, ranging from a low of perhaps 2 to 3 Mc. at night in a sunspot minimum to a high of 12 or 13 Mc. in daytime during a sunspot maximum. Whenever the critical frequency is above an amateur band, it is possible to communicate on that band over all distances from zero to the maximum that absorption will permit.

Maximum Usable Frequency

Of more interest, from a practical standpoint, than the critical frequency is the frequency *range* over which communication can be carried on via one or the other of the two reflecting layers.

In particular, it is useful to know the maximum usable frequency (abbreviated m.u.f.) for a particular distance at the time of day at which communication is desired. It is always advantageous to use the highest possible frequency because the absorption is less the higher the frequency. Therefore the m.u.f. always gives the greatest signal strength at the receiving point for a given transmitting power. The distance to be considered in determining the m.u.f. is the length of one hop in multihop transmission.

The m.u.f. depends upon the critical frequency and thus is subject to seasonal variations as well as variations throughout the day. To employ the m.u.f. for very long distances with the smallest number of hops requires that the antenna system radiate well at very low vertical angles.

As the frequency is decreased below the m.u.f., the signal strength also decreases because of greater absorption. Eventually, as the frequency continues to be lowered, the signal will disappear in the noise background that is always present. Thus there is a low-frequency limit, under a given set of ionosphere conditions, as well as a high-frequency limit to the range of frequencies that can be used for a given distance. The lowest useful high frequency (abbreviated l.u.h.f.) depends considerably on the transmitter power available, since high power will push the signal through the noise where low power would fail. But when the frequency is near the m.u.f., even low-power signals often will give surprising signal strength at long distances.

In commercial communication it is considered good practice to operate on a frequency about 15 per cent below the m.u.f. This allows for day-to-day variations in the ionosphere. This somewhat lower frequency is known as the optimum working frequency (abbreviated o.w.f.) or optimum traffic frequency (f.o.t.). Since amateur stations work in fixed bands of frequencies, it is not possible to choose either the m.u.f. or o.w.f. at will. Instead, the *time of day* at which optimum conditions can be expected for a given distance on a particular band must be determined.

Transmission Distance and Layer Height

Consideration of Fig. 1-8 shows that the distance at which a particular ray returns to earth depends upon the angle at which it enters the layer. This angle in turn is determined by the angle (called the wave angle) at which the ray leaves the transmitting antenna. Not shown in the drawing, but inherent in the geometry of the situation, is the fact that the distance also depends on the layer height. As the layer height is increased, the distance at which a ray leaving at a fixed wave angle returns to earth also increases. The same wave angle, therefore, will result in transmission over a greater distance when the wave is reflected by the F_2 layer than when it is reflected by the E layer.

The maximum distance that can be covered by one-hop transmission is approximately 1250 miles when the reflection is from the E layer, and approximately 2500 miles when the reflection is from the F_2 layer. These distances are based on average virtual heights, and in both cases a wave angle of zero (ray leaving the antenna tangent to the earth) is required. The actual distance covered by good one-hop transmission is somewhat less, at least at frequencies below 28 Mc., because of ground losses at wave angles below about 3 degrees.

The wave angle required for distances less than the maximum are shown in the chart of Fig. 1-10. The curves are based on average values of virtual height, and are for one-hop transmission. For two or more hops, the distance should be divided by the number of hops and the wave angle read from the chart on the basis of a single hop for the shorter distance. The critical angle

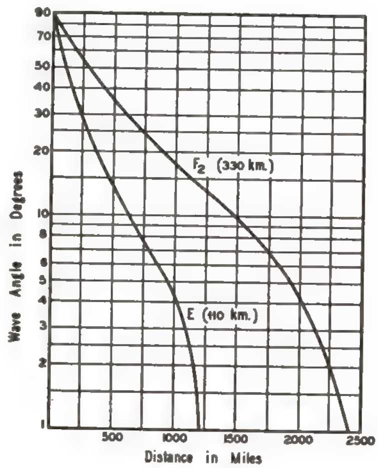


Fig. 1-10—Distance plotted against wave angle (one-hop transmission) for average virtual heights of the E and F_2 layers.

must, of course, be greater than the wave angle that is required for the number of hops selected.

One-hop transmission, when possible, will provide the greatest signal strength at the receiving point because there is some energy loss at each reflection, whether in the ionosphere or at the earth. At the longer distances, this requires a small wave angle, or "low-angle radiation" from the antenna. High-angle radiation is most useful for covering short distances. It will be appreciated that "long distance" and "short distance" are relative terms when it is remembered that the distance depends on the layer height as well as the wave angle. At times when the frequency in use is reflected by the *E* layer the distance will be one thing, but at another time of day when the *E* layer is ineffective and the F_2 layer comes into play the same wave angle from the same antenna will cover a much larger distance. That is one reason why it is possible to communicate over longer distances at night on frequencies in the vicinity of 7 Mc. than it is in the daytime.

Long-Distance Transmission

From the discussion in the preceding section, it should be clear that transmission over distances greater than 2500 miles must involve multihop propagation, because 2500 miles is the maximum distance that can be covered by one hop via the highest layer. Since multihop transmission increases the energy loss, it is quite important, for most effective long-distance transmission, that a frequency near the m.u.f. be used, and that the antenna concentrate the radiation at low angles so that the number of reflections will be small.

The propagation of waves over long paths is complicated by a number of factors. For example, at the particular frequency used the *E* layer may reflect the waves along part or parts of the path while the *F* layer does the reflecting at other parts. This will depend on the time of day, whether the path is generally north-south or east-west, the part of the world over which the path lies—in short, on the state of the ionosphere all along the path. It is also possible that a wave reflected downward from the F_2 layer will be reflected *upward* from the *E* layer instead of being reflected from the earth. However, all these possibilities have but little effect on the primary consideration in DX-antenna design—that the antenna should concentrate the radiation at the lowest possible angle.

Despite the complexity of long-distance propagation, there is a relatively simple method of determining average communication possibilities in advance. This is based on control points located 1250 miles from the transmitter and receiver, respectively, along the great-circle path connecting them. If the m.u.f. at the transmitter's control point is, say, 14 Mc., transmission in the direction of the receiver is possible on that frequency. If the m.u.f. at the receiver's control point is 14 Mc. or higher the signal will be heard. On the other hand, if the m.u.f. at the receiving control point is 10 Mc., a 14-Mc. signal from the transmitter will not be heard. The transmitting frequency must be lowered to 10 Mc. before communication is possible. In other words, the *lower* of the m.u.f.s at the two control points is the m.u.f. for the *circuit*. The values of m.u.f. at control points in any part of the world can be determined in advance from the CRPL charts mentioned on p. 23 in this chapter. While communication is possible, in theory, at any frequency below the circuit m.u.f., in practice the absorption becomes too great if the frequency is lowered too much below the m.u.f.

The 1250-mile control point is used for F_2 transmission. The control point is 625 miles away when the *E* layer is effective This may occur at either end of the circuit. If the frequency to be used is below the *E*-layer m.u.f. at that particular time, the *E* layer will control at the end of the circuit at which it is operating. This fact should not be forgotten when using the charts, because it frequently happens that the *E* layer is controlling at one end when the F_2 layer is controlling at the other. Under such circumstances the F_2 m.u.f. may be so high at both ends of the circuit that high absorption would be expected, whereas the actual case is that rather good signals will be received because the operating frequency is near the *E*-layer m.u.f. at one or both ends.

The control-point method of prediction does not explain *how* the waves travel from the transmitter to the receiver. Its justification is that it has been found to be a useful method, on the average, for predicting whether or not communication will be possible at a given frequency, or for selecting a frequency that will give communication between any two points. It does not give accurate results on east-west paths at distances just over the maximum one-hop range, since in this case the actual reflection points in the ionosphere are at a considerable distance from the selected control point. At greater distances, and especially where there are several possible numbers of hops, the system works well.

The vertical angle at which a wave arrives at the receiving point in long-distance transmission has been found by measurement to vary over a considerable range. For example, measurements on a path from England to the New Jersey coast indicate that on 7 Mc. the wave angle of the received signal at times is as high as 35 degrees, and on 14 Mc. is at times as high as 17 degrees. For 99 per cent of the time it is below those figures on these two frequencies. On the other hand, the same measurements showed that for 99 per cent of the time the angle was above 10 degrees on 7 Mc. and above 6 degrees on 14 Mc. For about half the time the angle was between 22 and 35 degrees on 7 Mc. and between 11 degrees and 17 degrees on 14 Mc. Whether or not there is exact reciprocity between the transmitting and receiving wave angles, these figures indicate the importance of keeping the wave angle low. They also show that the higher the frequency the less useful the higher wave angles become, in transmission over long distances.

Polarization and Direction of Travel

Because of the nature of refraction in the ionosphere, the polarization of the refracted wave usually is shifted from the direction it had on leaving the transmitting antenna. It is therefore not at all necessary to use antennas having the same polarization at the receiving and transmitting points. At the frequencies for which sky waves are useful, most amateurs use horizontal antennas. Depending on the type, such antennas may generate either horizontally- or elliptically-polarized waves.

For the most part, a wave follows the most direct path between the transmitter and receiver. In other words, it follows the great circle connecting the two points. Because of variations in the ionosphere the actual path may vary slightly, and shifts of as much as 5 degrees from the true great-circle path occur at times.

There are always two great-circle paths connecting two points on the earth's surface, one representing the shortest distance between them and the other a path in exactly the opposite direction—around the world the other way. Most communication is via the "short path." However, "long-path" communication is not uncommon, particularly when there is not too much difference between the two distances. At certain times of the day, when the short path would be inoperative, the ionosphere may be able to support communication over the long path.

Occasionally waves arrive from directions that seem to bear no visible relationship to the direction in which the transmitting station lies. While there are well-authenticated cases of this, and reasonable explanations have been worked out on the basis of known behavior of the ionosphere, it is probable that the apparent shift in direction frequently observed by amateurs is a result of "scattering," described in a later section. It is also possible that a combination of the vertical angle at which the wave arrives and minor responses of the antenna system used gives a false direction indication. Accurate direction finding with the sky wave at high frequencies is extremely difficult, requiring highly-specialized design and construction of equipment.

MISCELLANEOUS FEATURES OF SKY-WAVE PROPAGATION

Although not having any very direct bearing on antenna design, there are several aspects of sky-wave transmission that are of considerable interest from an operating standpoint. The ability to recognize and appraise unusual propagation effects often will help to explain seeming inconsistencies that may wrongly be blamed on faulty antenna design.

Ionosphere Variations

The daily and seasonal variations in the ionized layers that result from changes in the amount of ultraviolet light received from the sun have already been mentioned. Reference has also been made to the 11-year sunspot cycle, which directly affects propagation conditions because there is a rather direct correlation between sunspot activity and ionization. The 11-year figure for the time between successive peaks of sunspot activity is only an average; any given cycle may vary a few years either way. The last peak was during the winter of 1957-58. This peak and the one preceding it in 1947-48 were unusually high ones, as sunspot activity goes. On occasions the F_2 m.u.f. rose above 50 Mc. The cycle is considered to be almost at maximum as of the time this is being written (mid-1968). At a sunspot minimum, there is a period of a year or two when the F_2 m.u.f. does not get as high as 28 Mc. in temperate latitudes.

A small, but regular, variation in sunspot activity occurs over a period of about 28 days. This is the time required for the sun to make one rotation on its axis. The consequent rise and fall of

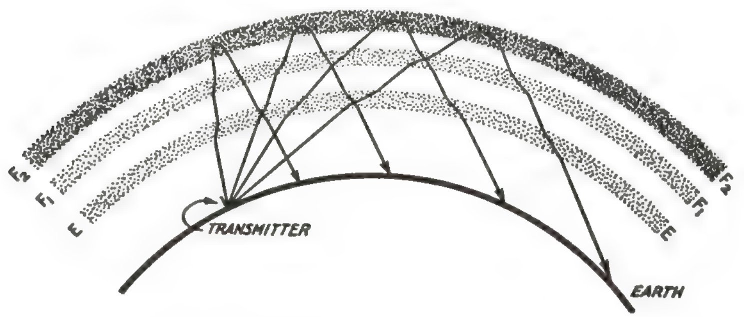


Fig. 1-11—Typical daytime propagation of high frequencies (14 to 28 Mc.). The waves are partially bent in going through the E and F_1 layers, but not enough to be returned to earth. The actual reflection is from the F_2 layer.

the m.u.f. makes a noticeable change in propagation conditions at frequencies from 14 to 28 Mc.

Fading

Variations in the strength of a received signal are classified under the general term **fading.** Long-period variations are to be expected through the day, on any given frequency, because the absorption changes with the height of the sun. There is also the daily variation of the m.u.f.; when the m.u.f. drops below the frequency in use the signal will "fade out."

In addition, the ionization at any part of the layer is in a continual state of change; there is turbulence in the ionosphere just as there is some turbulence in the atmosphere even on quiet days when the weather seems stable. The amount of absorption is continually varying; waves entering the ionosphere at slightly different angles will be refracted differently; the polarization is continually changing with refraction. The wave reaching the receiving antenna is usually made up of a group of rays each of which has been acted on a little differently by the ionosphere. Sometimes the rays are more or less in phase when they strike the receiving antenna; at other times some of the rays may be out of phase with others. The result is a continual variation in signal strength that may occur at rates varying from several times a second to once every few minutes.

When transmission conditions are not alike for waves of slightly different frequency, the sidebands in voice transmission may have a different fading pattern than the carrier. This is known as selective fading. It causes severe distortion of the modulation, especially when the carrier fades down while the sidebands do not. The distortion is in general worse with frequency modulation than with amplitude modulation, and is least with single-sideband carrierless transmission. Selective fading is more serious at the lower frequencies, such as 4 Mc., where the sideband frequencies represent a larger percentage of the carrier frequency than they do at a frequency such as 28 Mc.

Fading may be entirely different at two receiving points only a short distance apart. By the use of antennas separated by a wavelength or two, feeding separate receivers, it is possible to overcome the effects of amplitude fading, but not of selective fading. Similar use of inputs from antennas of differing polarization will often serve the same purpose. Such receiving arrangements are known as "diversity" systems.

Ionospheric Storms

Unusual eruptions on the sun cause disturbances in the ionosphere called ionospheric storms. These are accompanied by disturbances in the earth's magnetic field called magnetic storms. Storms of this type are most frequent during the sunspot cycle peak. They have a pronounced effect on radio communication.

The effect of an ionospheric storm on the lower frequencies is to decrease the daytime absorption and thus increase the daytime signal strength. At night the signal strength is below normal and is comparable to the daytime levels. On high frequencies communication frequently becomes impossible, as though the refracting layers had disappeared. The storms vary in intensity and duration. They may last from one to several days.

Ionospheric storms tend to recur at 28-day intervals since they are associated with particular sunspots or groups of sunspots, and these tend to maintain a more or less fixed position on the sun's surface as it rotates. The period of rotation, as mentioned before, is about 28 days.

Aurora

During magnetic storms auroral activity becomes more pronounced and extends farther from the polar regions than is normally the case. During abnormal auroral activity a peculiar form of wave propagation is frequently observed, in which the auroral curtain acts as a reflector. Waves directed toward the polar regions will be reflected back and can be used for communication on frequencies and over distances that normally would be skipped over. When this condition obtains it is necessary, when directive antennas are used, that both the transmitting and receiving antennas be directed toward the polar regions rather than along the great-circle path between the two stations.

Auroral propagation is particularly noticeable on the 28- and 50-Mc. bands. It is easily identified by a very rapid "flutter" fading that in effect modulates the received signal. On v.h.f. the flutter is frequently so pronounced that voice-modulated signals cannot be understood, even though the signals are quite strong. Under these conditions only c.w. signals will provide intelligible communication.

Sudden Fade-Outs

For some reason not completely understood, sky-wave transmission occasionally ceases abruptly and a radio "black-out" occurs, lasting sometimes for a few minutes and sometimes several hours. These sudden fade-outs are always associated with an unusual eruption on the sun, and affect only the part of the earth illuminated by the sun. Although the whole sky-wave spectrum may be affected, the more usual case is that the lower-frequency part is completely blacked out while long-distance propagation continues at the high-frequency end above about 25 Mc.

During one of these sudden fade-outs there is a marked decrease in noise level, since much of the received noise originates at distant points and arrives by way of the ionosphere. The only signals that can be heard during a true fade-out

are those from stations within the ground-wave range.

Sporadic-E Layer Refraction

In addition to the normal variations in ionization in the E layer as previously described, there are also "patches" of relatively intense ionization scattered throughout the layer. These are of varying intensity and size, usually are moving, and appear and disappear apparently at random. They may occur at any time of the day or night, and their cause is unknown.

Sporadic-E ionization has no critical frequency. That is, there is no well-defined frequency at which a wave striking the ionized region ceases to be returned to earth. Instead, the intensity of the returning signal simply drops off as the frequency is raised until eventually it is too weak to be usable.

Sporadic-E ionization is present all the time, although not at all places nor in sufficient intensity to provide regular communication on high frequencies, such as 28 and 50 Mc., where the normal E layer is not operative. However, it will often take part in propagation over long distances at 14 Mc. and below, accounting for communications at times and between points where the F_2 layer alone would not support it. In the 3.5- and 7-Mc. bands it is more of a factor than is generally realized.

At times the intensity of ionization in a sporadic-E patch is much higher than in the normal E layer at its best. When this occurs sky-wave communication sometimes becomes possible at frequencies as high as the 50-Mc. band over distances as short as 500 miles. Provided the patch is situated midway between two stations, communication is possible to the limit of one-hop E-layer transmission, or approximately 1250 miles. There have been instances of two-hop 50-Mc. communication between the east and west coasts of the United States, but such instances are relatively infrequent because *two* sporadic-E "clouds" have to be located at just the right spots between the two stations.

Sporadic-E transmission is more common on 28 than on 50 Mc. because it can be supported by a lesser degree of ionization on the lower frequency. In other respects the behavior on the two bands is the same, except that when the ionization is high enough to permit 50-Mc. communication, the skip distance will be shorter on 28 Mc. than on 50 Mc.

The comparison between 28 Mc. and 50 Mc. can also be carried to the relationship between 14 Mc. and 28 Mc.; sporadic-E is frequently observed on 14 Mc. when it is not operative on 28 Mc., but when it is operative on both bands the skip distance will be shorter on 14 Mc. than it is on 28 Mc.

Maximum utilization of sporadic-E ionization at 28 and 50 Mc. requires antennas having low-angle characteristics, because the ionization is not intense enough to return high-angle waves to earth.

Scatter Signals

When a skip zone exists it might be expected, from the simplified explanations of propagation given earlier, that no signals at all would be heard from stations too near to be reached by the sky wave and too far away for the ground wave to be heard. Actually, however, signals from these stations usually can be detected. The strength is low and the signal has a "fluttery" or "warbly" fade that is very characteristic. The signal results from "scattered" radiation that arrives at the receiving point from random directions and in random phase relationships.

Scattering may take place in the ionosphere or may result from repeated reflections between the ionosphere and earth. It may even occur as a result of repeated reflections between the two ionized layers. The signal heard in the skip zone may have traveled thousands of miles, even though the transmitter and receiver are separated by less than 100 miles. The principal cause, however, is that some of the wave energy is reflected back toward the transmitter when it strikes the earth after its first hop. With special equipment, the actual skip distance can be measured from such reflections.

Scattered reflections often will permit communication to be maintained for periods after the normal type of transmission would be expected to die out because of decreasing maximum usable frequency.

Meteor Trails

Meteors entering the upper atmosphere travel at such high speed that a large amount of energy is released when the meteor is slowed down by friction with the air. Part of this is used in ionizing the atmosphere along the path followed by the meteor. Even a very small meteor can ionize a region 50 or more feet in diameter and a mile or so long. Such a region is large enough to refract the shorter wavelengths back to earth. The ions quickly recombine, however, so the effect of a meteor usually lasts only a short time—from a fraction of a second to a few seconds, in the average case. It is long enough, though, to produce a "burst" of signal from stations not normally heard, or heard only weakly by scattered propagation.

Bursts caused by meteors can be observed at amateur frequencies from 14 Mc. through 144 Mc. During meteor showers the bursts are so frequent that it is sometimes possible to carry on continuous communication on 28 and 50 Mc. by that means. Just as in the case of sporadic-E patches, the ionized meteor trail must be midway between two stations if a burst is to be heard, and the two stations must be separated by enough distance so that the wave angle will be low enough to be refracted.

PREDICTION CHARTS

Regular observations of the ionosphere, and correlation with observed signals from various

distances over different paths, have made it possible to predict the median maximum usable frequencies to be expected over periods of several months. These predictions, for both the E and F_2 layers, are issued by the Environmental Science Services Administration each month in the form of charts showing the predicted median m.u.f. three months in advance. The charts, the "ITS Ionospheric Predictions," are available from the Superintendent of Documents, U. S. Government Printing Office, Washington, D. C. 20402, at 25 cents per copy, or on an annual subscription basis (12 issues) for $2.75 (foreign countries, $3.50).

The method of using the charts is described in *NHS Handbook 90*, available from the Supt. of Documents for 40 cents. A description of the charts and their uses is outside the scope of this book, but any amateur interested in communication over long distances will find them extremely valuable. They are not at all difficult to use.

Antenna Fundamentals

An antenna is an electric circuit of a special kind. In the ordinary type of circuit the dimensions of coils, capacitors and connections usually are small compared with the wavelength that corresponds to the frequency in use. When this is the case most of the electromagnetic energy stays in the circuit itself and either is used up in performing useful work or is converted into heat. But when the dimensions of wiring or components become appreciable compared with the wavelength some of the energy escapes by **radiation** in the form of electromagnetic waves. When the circuit is intentionally designed so that the major portion of the energy is radiated, we have an **antenna.**

Usually, the antenna is a straight section of conductor, or a combination of such conductors. Very frequently the conductor is a wire, although rods and tubing also are used. In this chapter we shall use the term "wire" to mean any type of conductor having a cross section that is small compared with its length.

The strength of the electromagnetic field radiated from a section of wire carrying radio-frequency current depends on the length of the wire and the amount of current flowing.* Other things being equal, the field strength will be directly proportional to the current. It is therefore desirable to make the current as large as possible, considering the power available. In any circuit that contains both resistance and reactance, the largest current will flow (for a given amount of power) when the reactance is "tuned out"—in other words, when the circuit is made **resonant** at the operating frequency. So it is with the common type of antenna; the current in it will be largest, and the radiation therefore greatest, when the antenna is resonant.

In an ordinary circuit the inductance is usually concentrated in a coil, the capacitance in a capacitor, and the resistance is principally concentrated in resistors, although some may be distributed around the circuit wiring and coil conductors. Such circuits are said to have **lumped constants.** In an antenna, on the other hand, the inductance capacitance and resistance are distributed along the wire. Such a circuit is said to have **distributed constants.** Circuits with distributed constants are so frequently straight-line conductors that they are customarily called linear circuits.

* It would also be true to say that the field strength depends on the voltage across the section of wire, but it is generally more convenient to measure current. The electromagnetic field consists of both magnetic and electric energy, with the total energy equally divided between the two. One cannot exist without the other in an electromagnetic wave, and the voltage in an antenna is just as much a measure of the field intensity as the current.

RESONANCE IN LINEAR CIRCUITS

The shortest length of wire that will resonate to a given frequency is one just long enough to permit an electric charge to travel from one end to the other and then back again in the time of one r.f. cycle. If the speed at which the charge travels is equal to the velocity of light, 300,000,000 meters per second, the distance it will cover in one cycle will be equal to this velocity divided by the frequency in cycles per second, or

$$\lambda = \frac{300,000,000}{f}$$

in which λ is the wavelength in meters. Since the charge traverses the wire *twice,* the length of wire needed to permit the charge to travel a distance λ in one cycle is $\lambda/2$, or one-half wavelength. Therefore the shortest *resonant* wire will be a half wavelength long.

The reason for this length can be made clear by a simple example. Imagine a trough with barriers at each end. If an elastic ball is started along the trough from one end, it will strike the far barrier, bounce back, travel along to the near barrier, bounce again, and continue until the energy imparted to it originally is all dissipated. If, however, whenever it returns to the near barrier it is given a new push just as it starts away, its back-and-forth motion can be kept up indefinitely. The impulses, however, must be *timed* properly; in other words, the rate or frequency of the impulses must be adjusted to the length of travel and the rate of travel. Or, if the timing of the impulses and the speed of the ball are fixed, the length of the trough must be adjusted to "fit."

In the case of the antenna, the speed is essentially constant, so we have the alternatives of adjusting the frequency to a given length of wire, or the length of wire to a given operating frequency. The latter is usually the practical condition.

By changing the units in the equation just given, and dividing by 2, the more useful formula is obtained.

$$l = \frac{492}{f\,(\text{Mc.})}$$

In this case l is the length *in feet* of a *half* wavelength for a frequency f, given in megacycles, when the wave travels with the velocity of light. This formula is the basis upon which several significant lengths in antenna work are developed. It represents the length of a half wavelength in space, when no factors that modify the speed of propagation exist.

Current and Voltage Distribution

If the wire in the first illustration had been infinitely long the charge (voltage) and the current (an electric current is simply a charge in motion) would both slowly decrease in amplitude with distance from the source. The slow decrease would result from dissipation of energy in the form of radio waves and in heating the wire because of its resistance. However, when the wire is short the charge is reflected when it reaches the far end, just as the ball bounced back from the barrier. With radio-frequency excitation of a half-wave antenna, there is of course not just a single charge but a continuous supply of energy, varying in voltage according to a sine-wave cycle. We might consider this as a series of charges, each of slightly different amplitude than the preceding one. When a charge reaches the end of the antenna and is reflected, the direction of current flow reverses, since the charge is now traveling in the opposite direction. However, the next charge is just reaching the end of the antenna, so we have two currents of practically the same amplitude flowing in opposite directions. The resultant current at the end of the antenna therefore is zero. As we move farther back from the end of the antenna the magnitudes of the outgoing and returning currents are no longer the same because the charges causing them have been supplied to the antenna at different parts of the r.f. cycle. There is less cancellation, therefore, and a measurable current exists. The greatest difference—that is, the largest resultant current—will be found to exist a quarter wavelength away from the end of the antenna. As we move back still farther from this point the current will decrease until, a half wavelength away from the end of the antenna, it will reach zero again. Thus in a half-wave antenna the current is zero at the ends and maximum at the center.

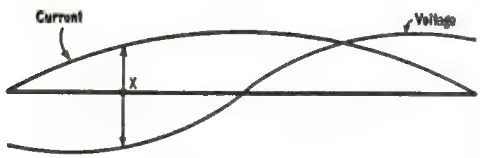


Fig. 2-1—Current and voltage distribution on a half-wave wire. In this conventional representation the distance at any point (X, for instance) from the wire, represented by the heavy line, to the curve gives the relative intensity of current or voltage at that point. The relative direction of current flow (or polarity of voltage) is indicated by drawing the curve either above or below the line that represents the antenna. The curve above, for example, shows that the instantaneous polarity of the voltage in one half of the antenna is opposite to that in the other half.

This current distribution along a half-wave wire is shown in Fig. 2-1. The distance measured vertically from the antenna wire to the curve marked "current," at any point along the wire, represents the relative amplitude of the current as measured by an ammeter at that point. This is called a **standing wave** of current. The *instantaneous* value of current at any point varies sinusoidally at the applied frequency, but its amplitude is different at every point along the wire as shown by the curve. The standing-wave curve itself has the shape of a half sine wave, at least to a good approximation.

The voltage along the wire will behave differently; it is obviously greatest at the end since at this point we have two practically equal charges adding. As we move back along the wire, however, the outgoing and returning charges are not equal and their sum is smaller. At the quarter-wave point the returning charge is of equal magnitude but of opposite sign to the outgoing charge, since at this time the polarity of the voltage wave from the source has reversed (one-half cycle). The two voltages therefore cancel each other and the resultant voltage is zero. Beyond the quarter-wave point, away from the end of the wire, the voltage again increases, but this time with the opposite polarity.

It will be observed, therefore, that the voltage is maximum at every point where the current is minimum, and vice versa. The polarity of the current or voltage reverses every half wavelength along the wire, but the reversals do not occur at the same points for both current and voltage; the respective reversals occur, in fact, at points a quarter wave apart.

A maximum point on a standing wave is called a loop (or antinode); a minimum point is called a node.

Harmonic Operation

If there is reflection from the end of a wire, the number of standing waves on the wire will be equal to the length of the wire divided by a half wavelength. Thus, if the wire is two half waves long there will be two standing waves; if three half waves long, three standing waves, and so on. These longer wires, each multiples of a half wave in length, also will be resonant, therefore, to the same frequency as the single half-wave wire. When an antenna is two or more half waves in length at the operating frequency it is said to be **harmonically resonant,** or to operate at a harmonic. The number of the harmonic is

the number of standing waves on the wire. For example, a wire two half-waves long is said to be operating on its **second harmonic**; one three half-waves long on its **third harmonic**, and so on.

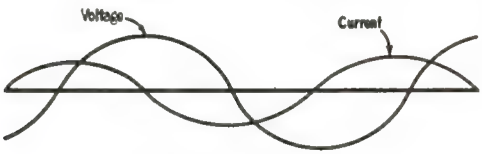


Fig. 2-2—Harmonic operation of a long wire. The wire is long enough to contain several half waves. The current and voltage curves cross the heavy line representing the wire to indicate that there is reversal in the direction of the current, and a reversal in the polarity of the voltage, at intervals of a half wavelength. The reversals of current and voltage do not coincide, but occur at points a quarter wavelength apart.

Harmonic operation is often utilized in antenna work because it permits operating the same antenna on several harmonically-related amateur bands. It is also an important principle in the operation of certain types of directive antennas.

Electrical Length

The *electrical* length of a linear circuit such as an antenna wire is not necessarily the same as its *physical* length in wavelengths or fractions of a wavelength. Rather, the electrical length is measured by the *time* taken for the completion of a specified phenomenon.

For instance, we might imagine two linear circuits having such different characteristics that the speed at which a charge travels is not the same in both. Suppose we wish to make both circuits resonant to the same frequency, and for that purpose adjust the physical length of each until a charge started at one end travels to the far end, is reflected and completes its return journey to the near end in exactly the time of one r.f. cycle. Then it will be found that the physical length of the circuit with the lower velocity of propagation is shorter than the *physical* length of the other. The *electrical* lengths, however, are identical, each being a half-wavelength.

In alternating-current circuits the instantaneous values of current or voltage are determined by the instant during the cycle at which the measurement is made (assuming, of course, that such a measurement could be made rapidly enough). If the current and voltage follow a sine curve—which is the usual case—the time, for any instantaneous value, can be specified in terms of an angle. The sine of the angle gives the instantaneous value when multiplied by the *peak* value of the current or voltage. A complete sine curve occupies the 360 degrees of a circle and represents one cycle of a.c. current or voltage. Thus a half cycle is equal to 180 degrees, a quarter cycle to 90 degrees, and so on.

It is often convenient to use this same form of representation for linear circuits. When the electrical length of such a circuit is such that a charge, *traveling in one direction,* takes the time of one cycle to traverse it, the length of the circuit is said to be 360 degrees. This corresponds to one wavelength. On a wire a half-wave in electrical length the charge completes a one-way journey in one-half cycle and its length is said to be 180 degrees. The angular method of measurement is quite useful for lengths that are not easily-remembered fractions or simple multiples of such fractions. A chart for converting fractions of a wavelength to degrees is given in Fig. 2-3.

Velocity of Propagation

The velocity at which electromagnetic waves travel through a medium depends upon the dielectric constant of the medium. At r.f. the dielectric constant of air is practically unity, so the waves travel at essentially the same velocity as light in a vacuum. This is also the velocity, very closely, of the charge traveling along a wire.

If the dielectric constant is greater than 1, the velocity of propagation is lowered. Thus the introduction, in appreciable quantity, of insulating material having a dielectric constant greater than 1 will cause the wave to slow down. This effect is encountered in practice in connection with both antennas and transmission lines. It causes the electrical length of the line or antenna to be greater than actual physical length.

Length of a "Half-Wave" Antenna

Even if the antenna could be supported by insulators that did not cause the electromagnetic fields traveling along the wire to slow down, the physical length of a practical antenna always is somewhat less than its electrical length. That is, a "half-wave" antenna is not one having the

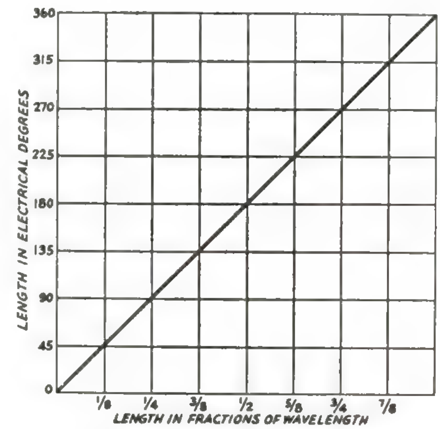


Fig. 2-3—Chart for converting electrical degrees to fractions of a wavelength.

same length as a half wavelength in space. It is one having an *electrical* length equal to 180 degrees. Or, to put it another way, it is one whose length has been adjusted to "tune out" any reactance, so it is a *resonant* antenna.

The antenna length required to resonate at a given frequency (independently of any dielectric effects) depends on the ratio of the length of the conductor to its diameter. The smaller this ratio, the shorter the antenna for a given electrical length. This effect is shown in Fig. 2-4 as a factor (K) by which a free-space half wavelength must be multiplied to find the resonant length, as a function of the ratio of the free-space half wavelength to conductor diameter, known as the **length/diameter ratio.** The curve is based on theoretical considerations and is useful as a guide to the probable antenna length for a given frequency. It applies only to conductors of uniform diameter (tapered elements such as are used in some types of beam antennas will generally be longer, for the same frequency) and does not include any effects introduced by the method of supporting the conductor.

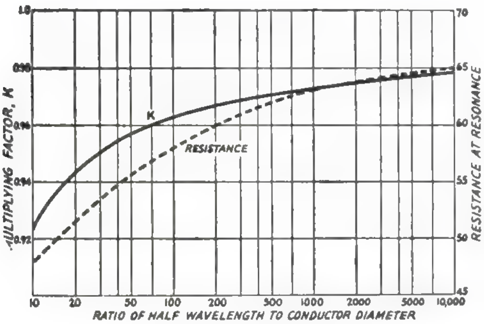


Fig. 2-4—The solid curve shows the factor, *K*, by which the length of a half wave in free space should be multiplied to obtain the physical length of a resonant half-wave antenna having the length/diameter ratio shown along the horizontal axis. The broken curve shows how the radiation resistance of a half-wave antenna varies with the length/diameter ratio.

A length/diameter ratio of 10,000 is roughly average for wire antennas (actually, it is approximately the ratio for a 7-Mc. half-wave antenna made of No. 12 wire). In this region K changes rather slowly and a half-wave antenna made of wire is about 2 per cent shorter than a half wavelength in space.

The shortening effect is most pronounced when the length/diameter ratio is 100 or less. An antenna constructed of 1-inch diameter tubing for use on 144 Mc., for example, would have a length/diameter ratio of about 40 and would be almost 5 per cent shorter than a free-space half wavelength.

If the antenna is made of rod or tubing and is not supported near the ends by insulators, the following formula will give the required physical length of a half-wave antenna based on Fig. 2-4.

$$\textit{Length}\ (\text{feet}) = \frac{492 \times K}{f\ (\text{Mc.})}$$

or

$$\textit{Length}\ (\text{inches}) = \frac{5905 \times K}{f\ (\text{Mc.})}$$

where K is taken from Fig. 2-4 for the particular length/diameter ratio of the conductor used.

End Effect

If the formulas of the preceding section are used to determine the length of a wire antenna, the antenna will resonate at a somewhat lower frequency than is desired. The reason for this is that there is an additional "loading" effect caused by the insulators that must be used at the ends of the wire for suspending it. These insulators and the wire loop that ties the insulator to the antenna add a small amount of capacitance to the system. This capacitance helps to tune the antenna to a slightly lower frequency, in much the same way that additional capacitance in any tuned circuit will lower the resonant frequency. In an antenna it is called **end effect.** The current at the ends of the antenna does not quite reach zero because of the end effect, as there is some current flowing into the end capacitance.

End effect increases with frequency and varies slightly with different installations. However, at frequencies up to 30 Mc. (the frequency range over which wire antennas are most likely to be used) experience shows that the length of a half-wave antenna is of the order of 5 per cent less than the length of a half wave in space. As an average, then, the physical length of a resonant half-wave antenna may be taken to be

$$l\ (\text{feet}) = \frac{492 \times 0.95}{f\ (\text{Mc.})}$$

or

$$l\ (\text{feet}) = \frac{468}{f\ (\text{Mc.})}$$

This formula is reasonably accurate for finding the physical length of a half-wave antenna for a given frequency, but does not apply to antennas longer than a half wave in length. In the practical case, if the antenna length must be adjusted to exact frequency (not all antenna systems require it) the length should be "pruned" to resonance.

ANTENNA IMPEDANCE

In the simplified description given earlier of voltage and current distribution along an antenna it was stated that the voltage was zero at the center of a half-wave antenna (or at any current loop along a longer antenna). It is more accurate to say that the voltage reaches a *minimum* rather than zero. Zero voltage with a finite value of current would imply that the circuit is

entirely without resistance. It would also imply that no energy is radiated by the antenna, since a circuit without resistance would take no real power from the driving source.

Actually, of course, an antenna, like any other circuit, consumes power. The current that flows in it therefore must be supplied at a finite value of voltage. The **impedance** of the antenna is simply equal to the voltage applied to its terminals divided by the current flowing into those terminals. If the current and voltage are exactly in phase the impedance is purely resistive. This is the case when the antenna is resonant. If the antenna is not exactly resonant the current will be somewhat out of phase with the applied voltage and the antenna shows reactance along with resistance.

Most amateur transmitting antennas are operated at or quite close to resonance so that reactive effects are in general comparatively small. They are nevertheless present, and must be taken into account, whenever an antenna is operated at other than the exact design frequency.

In the following discussion it is assumed that power is applied to the antenna by opening the conductor at the center and applying the driving voltage across the gap. This is shown in Fig. 2-5. While it is possible to supply power to the antenna by other methods, the selection of different driving points leads to different values of impedance; this can be appreciated after study of Fig. 2-1, which shows that the ratio of voltage to current (which is, by definition, the impedance) is different at every point along the antenna. To avoid confusion it is desirable to use the conditions at the center of the antenna as a basis.

The Antenna as a Circuit

If the frequency applied at the center of a half-wave antenna is varied above and below the resonant frequency, the antenna will exhibit much the same characteristics as a conventional series-resonant circuit. Exactly at resonance the current at the input terminals will be in phase with the applied voltage. If the frequency is on the low side of resonance the phase of the current will lead the voltage; that is, the reactance of the antenna is capacitive. When the frequency is on the high side of resonance the opposite occurs; the current lags the applied voltage and the antenna exhibits inductive reactance.

It is not hard to see why this is so. Consider the antennas shown in Fig. 2-6, one resonant, one too long for the applied frequency, and one

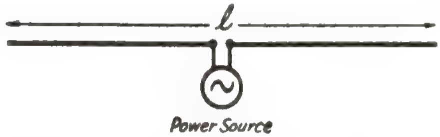


Fig. 2-5—The center-fed antenna discussed in the text. It is assumed that the leads from the source of power to the antenna have zero length.

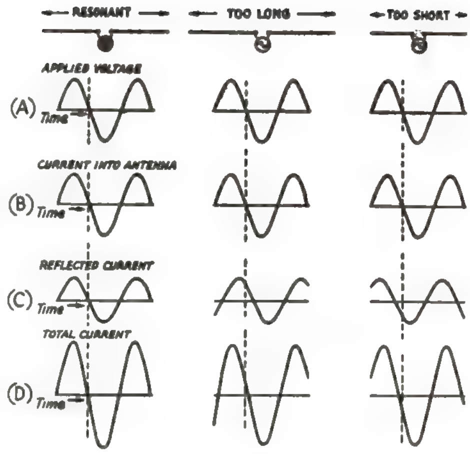


Fig. 2-6—Current flow in resonant and off-resonant antennas. The initial current flow, B, caused by the source of power, is in phase with the applied voltage, A. This is the outgoing current discussed in the text. The reflected current, C, combines with the outgoing current to form the resultant current, D, at the input terminals of the antenna.

too short. In each case the applied voltage is shown at A, and the instantaneous current going *into* the antenna because of the applied voltage is shown at B. Note that this current is always in phase with the applied voltage, regardless of the antenna length. For the sake of simplicity only the current flowing in one leg of the antenna is considered; conditions in the other leg are similar.

In the case of the resonant antenna, the current travels out to the end and back to the driving point in one-half cycle, since one leg of the antenna is 90 degrees long and the total path out and back is therefore 180 degrees. This would make the phase of the *reflected* component of current differ from that of the outgoing current by 180 degrees, since the latter current has gone through a half cycle in the meantime. However, it will be remembered that there is a phase shift of 180 degrees at the end of the antenna, because the direction of current reverses at the end. The *total* phase shift between the outgoing and reflected currents, therefore, is 360 degrees. In other words, the reflected component arrives at the driving point exactly in phase with the outgoing component. The reflected component, shown at C, adds to the outgoing component to form the resultant or total current at the driving point. The resultant current is shown at D, and in the case of the resonant antenna it is easily seen that the resultant is exactly in phase with the applied voltage. This being the case, the load seen by the source of power is a pure resistance.

Now consider the antenna that is too short to be resonant. The outgoing component of current is still in phase with the applied voltage, as

shown at B. The reflected component, however, gets back to the driving point *too soon,* because it travels over a path less than 180 degrees, out and back. This means that the maximum value of the reflected component occurs at the driving point ahead of (in time) the maximum value of the outgoing component, since that particular charge took less than a half cycle to get back. Including the 180-degree reversal at the end of the antenna, the total phase shift is therefore less than 360 degrees. This is shown at C, and the resultant current is the combination of the outgoing and reflected components as given at D. It can be seen that the resultant current leads the applied voltage, so the antenna looks like a resistance in series with a capacitance. The shorter the antenna the greater the phase shift between voltage and current; that is, the capacitive reactance increases as the antenna is shortened.

When the antenna is too long for the applied frequency the reflected component of current arrives too late to be exactly in phase with the outgoing component, because it must travel over a path more than 180 degrees long. The maximum value of the reflected component therefore occurs later (in time) than the maximum value of the outgoing component, as shown at C. The resultant current at the antenna input terminals therefore lags behind the applied voltage. The phase lag increases as the antenna is made longer. That is, an antenna that is too long shows inductive reactance along with resistance, and this reactance increases with an increase in antenna length over the length required for resonance.

If the antenna length is increased to 180 degrees on each leg the go-and-return path length for the current becomes 360 degrees. This, plus the 180-degree reversal at the end, makes the total phase shift 540 degrees, which is the same as a 180-degree shift. In this case the reflected current arrives at the input terminals exactly *out* of phase with the outgoing component, so the resultant current is very small. The resultant is in phase with the applied voltage, so the antenna impedance is again purely resistive. The resistance under these conditions is very high, and the antenna has the characteristics of a parallel-resonant circuit. A *voltage* loop, instead of a *current* loop, appears at the input terminals when each leg of the antenna is 180 degrees long.

The amplitude of the reflected component is less than that of the component of current going into the antenna. This is the result of energy loss by radiation as the current travels along the wire. It is perhaps easier to understand if, instead of thinking of the electromagnetic fields as being brought into being by the current flow, we adopt the more fundamental viewpoint that *current flow along a conductor is caused by a moving electro-magnetic field.* When some of the energy escapes from the system because the field travels out into space, it is not hard to understand why the current becomes less the farther it travels. There is simply less energy left to cause it. The difference between the outgoing- and reflected-current amplitudes accounts for the fact that the current does not go to zero at a voltage loop, and a similar difference between the applied- and reflected-voltage components explains why the voltage does not go to zero at a current loop.

Resistance

The energy supplied to an antenna is dissipated in the form of radio waves and in heat losses in the wire and near-by dielectrics. The radiated energy is the useful part, but so far as the antenna is concerned it represents a loss just as much as the energy lost in heating the wire is a loss. In either case the dissipated power is equal to I^2R: in the case of heat losses, R is a real resistance, but in the case of radiation R is an assumed resistance, which, if present, would dissipate the power that actually disappears by radiation. This fictitious resistance is called the radiation resistance. The total power loss in the antenna is therefore equal to $I^2(R_0 + R)$, where R_0 is the radiation resistance and R the real resistance, or ohmic resistance.

In the ordinary half-wave antenna operated at amateur frequencies the power lost as heat in the conductor does not exceed a few per cent of the total power supplied to the antenna. This is because the r.f. resistance of copper wire even as small as No. 14 is very low compared with the radiation resistance of an antenna that is reasonably clear of surrounding objects and is not too close to the ground. Therefore it can be assumed that the ohmic loss in a reasonably well-located antenna is negligible, and that all of the resistance shown by the antenna is radiation resistance. As a radiator of electromagnetic waves such an antenna is a highly-efficient device.

The value of radiation resistance, as measured at the center of a half-wave antenna, depends on a number of factors. One is the location of the antenna with respect to other objects, particularly the earth. Another is the length/diameter ratio of the conductor used. In "free space"—with the antenna remote from everything else—the radiation resistance of a resonant antenna made of an infinitely-thin conductor is approximately 73 ohms. While there is no such thing as having an antenna in free space, it is a convenient basis for calculation because the modifying effect of the ground can be taken into account separately. If the antenna is at least several wavelengths away from ground and other objects, it can be considered to be in free space insofar as its own electrical properties are concerned. This condition can be met with antennas in the v.h.f. and u.h.f. range.

The way in which the free-space radiation resistance varies with the length/diameter ratio of a half-wave antenna is shown by the broken curve in Fig. 2-4. As the antenna is made thick-

er the radiation resistance decreases. For most wire antennas it is close to 65 ohms. It will usually lie between 55 and 60 ohms for antennas constructed of rod or tubing.

The actual value of the radiation resistance—at least so long as it is 50 ohms or more—has no appreciable effect on the radiation efficiency of the antenna. This is because the ohmic resistance is only of the order of 1 ohm with the conductors used for thick antennas. The ohmic resistance does not become important until the radiation resistance drops to very low values—say less than 10 ohms—as may be the case when several antennas are coupled together.

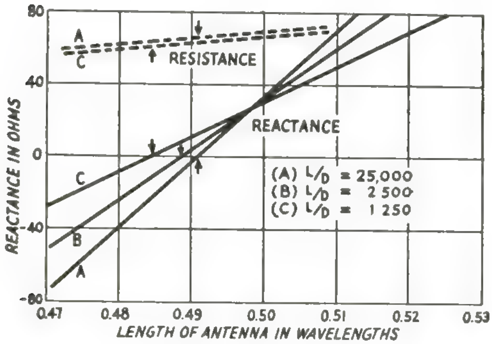


Fig. 2-7—Resistance and reactance at input terminals of a center-fed antenna as a function of its length near a half wavelength. As shown by curves A, B and C, the reactance is affected more by the length/diameter ratio of the conductor than is the radiation resistance.

The radiation resistance of a resonant antenna is the "load" for the transmitter or for the r.f. transmission line connecting the transmitter and antenna. Its value is important, therefore, in determining the way in which the antenna and transmitter or line are coupled together. The power must be supplied at the right voltage if the system as a whole is to be efficient.

The resistance of an antenna varies with its length as well as with the ratio of its length to its diameter. When the antenna is approximately a half wave long, the resistance changes rather slowly with length. This is shown by the curves of Fig. 2-7, where the change in resistance as the length is varied a few per cent on either side of resonance is shown by the broken curves. The resistance decreases somewhat when the antenna is slightly short, and increases when it is slightly long. These curves also illustrate the effect of changing the frequency applied to an antenna of fixed length, since increasing the frequency above resonance is the same thing as having an antenna that is too long, and vice versa.

The range covered by the curves in Fig. 2-7 is representative of the frequency range over which a fixed antenna is operated between the limits of an amateur band. At greater departures from the resonant length the resistance continues to decrease about uniformly as the antenna is shortened, but tends to increase rapidly as the antenna is made longer. The resistance increases very rapidly when the length of a leg exceeds about 135 degrees, or about ⅜ wavelength, and reaches a maximum when the length of one side is 180 degrees. This is considered in more detail in a later chapter.

Reactance

The rate at which the reactance of the antenna increases as the length is varied from resonance depends on the length/diameter ratio of the conductor. The thicker the conductor the smaller the reactance change for a given change in length. This is shown by the reactance curves in Fig. 2-7. Curves for three values of length/diameter ratio are shown; L represents the length of a half wave in space, approximately, and D is the diameter of the conductor in the same units as the length. The point where each curve crosses the zero axis (indicated by an arrow in each case) is the length at which an antenna of that particular length/diameter ratio is resonant. The effect of L/D on the resonant length also is illustrated by these curves: the smaller the ratio, the shorter the length at which the reactance is zero.

It will be observed that the reactance changes about twice as rapidly in the antenna with the smallest diameter (A) as it does in the antenna with the largest diameter (C). With still larger diameters the rate at which the reactance changes would be even smaller. As a practical matter it is advantageous to keep the reactance change with a given change in length as small as possible. This means that the reactance is comparatively low, when the antenna is operated over a small band of frequencies centered on its resonant frequency, and that the impedance change with frequency is small. This simplifies the problem of supplying power to the antenna when it must be worked at frequencies somewhat different from its resonant frequency.

At lengths considerably different from the resonant length the reactance changes more rapidly than in the curves shown in Fig. 2-7. As in the case of the resistance change, the change is most rapid when the length exceeds 135 degrees (⅜ wavelength) and approaches 180 degrees (½ wavelength) on a side. In this case the reactance is inductive and reaches a maximum at a length somewhat less than 180 degrees. Between this maximum point and 180 degrees of electrical length the reactance decreases very rapidly, becoming zero when the length is such as to be parallel-resonant.

Very short antennas have a large capacitive reactance. It was pointed out in the preceding section that with antennas shorter than 90 degrees on a side the resistance decreases at a fairly uniform rate, but this is not true of the reactance. It increases rather rapidly when the

length of a side is shortened below about 45 degrees.

The behavior of antennas with different length/diameter ratios corresponds to the behavior of ordinary resonant circuits having different *Q*s. When the *Q* of a circuit is low, the reactance is small and changes rather slowly as the applied frequency is varied on either side of the resonant frequency. If the *Q* is high, the converse is true. The response curve of the low-*Q* circuit is "broad"; that of the high-*Q* circuit "sharp". So it is with antennas; a thick antenna works well over a comparatively wide band of frequencies while a thin antenna is rather sharp in tuning. The *Q* of the thick antenna is low; the *Q* of the thin antenna is high, assuming essentially the same value of radiation resistance in both cases.

Coupled Antennas

A conventional tuned circuit far enough away from all other circuits so that no external coupling exists can be likened to an antenna in free space, in one sense. That is, its characteristics are unaffected by its surroundings. It will have a *Q* and resonant impedance determined by the inductance, capacitance and resistance of which it is composed, and those quantities alone. But as soon as it is coupled to another circuit its *Q* and impedance will change, depending on the characteristics of the other circuit and the degree of coupling.

A similar situation arises when two or more "elementary" antennas—half-wave antennas, frequently called **half-wave dipoles**—are coupled together. This coupling takes place merely by having the two antennas in proximity to each other. The sharpness of resonance and the radiation resistance of each "element" of the system are affected by the mutual interchange of energy between the coupled elements. The exact effect depends on the degree of coupling (that is, how close the antennas are to each other in terms of wavelength, and whether the wires are parallel or not) and the tuning condition (whether tuned to resonance or slightly off resonance) of each element. Analysis is extremely difficult and even then has to be based on some simplifying assumptions that may not be true in practice. Only a few relatively simple cases have been analyzed. Such data as are available for even moderately complicated systems of coupled antennas are confined to a few types and are based on experimental measurements. They are, therefore, subject to the inaccuracies that accompany any measurements in a field where measurement is difficult at best.

Antenna systems consisting of coupled elements will be taken up in later chapters. At this point it is sufficient to appreciate that the free-space values that have been discussed in this chapter may be modified drastically when more than one antenna element is involved in the system. It has already been pointed out that the presence of the ground, as well as nearby conductors or dielectrics, also will modify the free-space values. The free-space characteristics of the elementary half-wave dipole are only the point of departure for a practical antenna system. In other words, they give the basis for understanding antenna principles but cannot be applied too literally in the practical case.

It is of interest to note that the comparison between an isolated tuned circuit and an antenna in free space is likewise not to be taken too literally. In one sense the comparison is wholly misleading. The tuned circuit is usually so small, physically, in comparison with the wavelength, that practically no energy escapes from it by radiation. An antenna, to be worthy of the name, is always so large in comparison with the wavelength that practically *all* the energy supplied to it escapes by radiation. Thus the antenna can be said to be very tightly coupled to space, while the tuned circuit is not coupled to anything. This very fundamental difference is one reason why antenna systems cannot be analyzed as readily, and with as satisfactory results in the shape of simple formulas, as ordinary electrical circuits.

HARMONICALLY-OPERATED ANTENNAS

An antenna operated at a harmonic of its fundamental frequency has considerably different properties than the half-wave dipole previously discussed. It must first be emphasized that harmonic operation implies that there is a reversal of the direction of current flow in alternate half-wave sections of the antenna, as shown in Fig. 2-2 and again at A in Fig. 2-8. In the latter figure, the curve shows the standing wave of current intensity along the wire; the curve is above the line to indicate current flow in one direction (assumed to be to the right, in the direction of the arrow) and below the line to indicate current flow in the opposite direction in the other half-wave section. (During the next radio-frequency half cycle the current flow in the left-hand half-wave section would be toward the left, and in the right-hand half-wave section to the right; this alternation in direction takes place in each succeeding half cycle. However, the direction of current flow in adjacent half-wave sections is at all times opposite.) The antenna in this drawing is one wavelength long and is operating on its second harmonic.

Now consider the half-wave antenna shown at Fig. 2-8B. It is opened in the center and fed by a source of r.f. power through leads that are assumed to have zero length. Since one terminal of the generator is positive at the same instant that the other terminal is negative, current flows *into* one side of the generator while it is flowing *out* at the other terminal. Consequently the current flows in the same direction in both sections of the half-wave antenna. It has the amplitude distribution shown by the curve over the antenna wire.

If we now increase the length of the wire on

each side of the generator in Fig. 2-8B to one-half wavelength, we have the situation shown in Fig. 2-8C. At the instant shown, current flows into the generator from the left-hand half-wave section, and out of the generator into the right-hand half-wave section. Thus the currents in the two sections are in the same direction, just as they were in Fig. 2-8B. The current distribution in this case obviously is not the same as in Fig. 2-8A. Although the over-all lengths of the antennas shown at A and C are the same, the antenna at A is operating on a harmonic but the one in C is not.

For true harmonic operation it is necessary that the power be fed into the antenna at the proper point. Two methods that result in the proper current distribution are shown at D and E in Fig. 2-8. If the source of power is connected to the antenna at one end, as in D, the direction of current flow will be reversed in alternate half-wave sections. Or if the power is inserted at the center of a half-wave section, as in E, there will be a similar reversal of current in the next half-wave section. For harmonic operation, therefore, the antenna should be fed either at the end or at a current loop. If the feed point is at a current node the current distribution will not be that expected on a harmonic antenna.

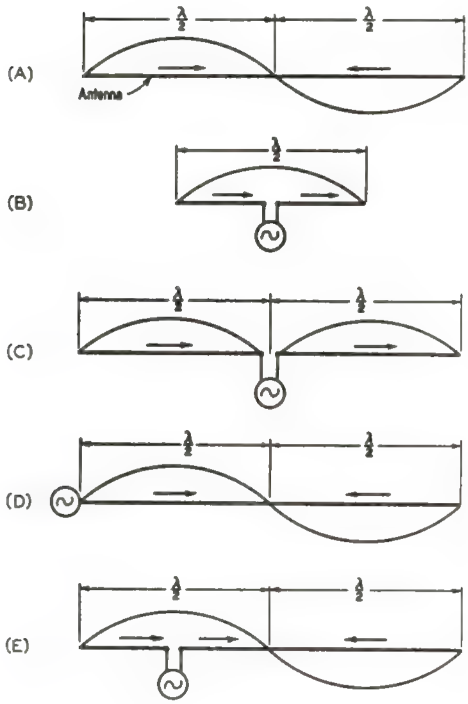


Fig. 2-8—How the feed point makes a difference in current distribution along the antenna. With center feed, increasing the length of each side of the antenna keeps the current flowing in the same direction in the two halves, up to the point where each side is a half wavelength long. For harmonic operation, the antenna must be fed in such a way that the current direction reverses in alternate half-wavelength sections. Suitable methods are shown at D and E.

Length of a Harmonic Wire

The physical length of a harmonic antenna is not exactly the same as its electrical length, for the same reasons discussed earlier in connection with the half-wave antenna. The physical length is somewhat shorter than the length of the same number of half waves in space because of the length/diameter ratio of the conductor and the end effects. Since the latter are appreciable only where insulators introduce additional capacitance at a high-voltage point along the wire, and since a harmonic antenna usually has such insulation only at the ends, the end-effect shortening affects only the half-wave sections at each end of the antenna. It has been found that the following formula for the length of a harmonic antenna of the usual wire sizes works out very well in practice:

$$\textit{Length}\ (\text{feet}) = \frac{492\ (N-0.05)}{\textit{Freq.}\ (\text{Mc.})}$$

where N is the number of *half* waves on the antenna.

Because the number of half waves varies with the harmonic on which the antenna is operated, consideration of this formula together with that for the half-wave antenna (the fundamental frequency) will show that the relationship between the antenna's fundamental frequency and its harmonics is not exactly integral. That is, the "second-harmonic" frequency to which a given length of wire is resonant is not exactly twice its fundamental frequency; the "third-harmonic" resonance is not exactly three times its fundamental, and so on. The actual resonant frequency on a harmonic always is a little higher than the exact multiple of the fundamental. A full-wave (second-harmonic) antenna, for example, must be a little longer than twice the length of a half-wave antenna.

Impedance of Harmonic Antennas

A harmonic antenna can be looked upon as a series of half-wave sections placed end to end (collinear) and supplied with power in such a way that the currents in alternate sections are out of phase. There is a certain amount of coupling between adjacent half-wave sections. Because of this coupling, as well as the effect of radiation from the additional sections, the impedance as measured at a current loop in a half-wave section is not the same as the impedance at the center of a half-wave antenna.

Just as in the case of a half-wave antenna, the impedance consists of two main components, radiation resistance and reactance. The ohmic or loss resistance is low enough to be ignored in the practical case. If the antenna is exactly resonant there will be no reactance at the input terminals and the impedance consists only of the radiation resistance. The value of the radiation resistance depends on the number of half waves on the wire and, as in the case of the half-wave antenna, is modified by the presence of near-by conductors and dielectrics, particularly the earth. As a point of departure, however, it is of interest to know the order of magnitude of the radiation resistance of a theoretical harmonic antenna consisting of an infinitely-thin conductor in free space, with its length adjusted to exact harmonic resonance. The radiation resistance of such an antenna having a length of one wavelength is

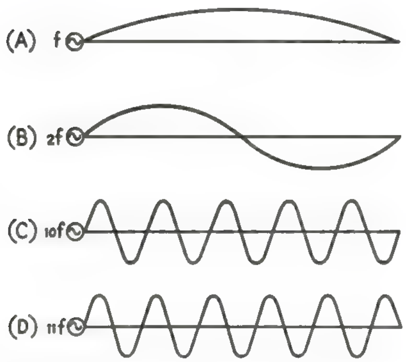


Fig. 2-9—The percentage frequency change from one high-order harmonic to the next (for example, between the 10th and 11th harmonics shown at C and D) is much smaller than between the fundamental and second harmonic (A and B). This makes impedance variations more rapid as the wire becomes longer in terms of wavelength.

approximately 90 ohms, and as the antenna length is increased the resistance also increases. At ten wavelengths it is approximately 160 ohms, for example. The way in which the radiation resistance of a theoretical harmonic wire varies with length is shown by curve A in Fig. 2-23, page 44. It is to be understood that the radiation resistance always is measured at a current loop.

When the antenna is operated at a frequency slightly off its exact resonant frequency, reactance as well as resistance will appear at its input terminals. In a general way, the reactance varies with applied frequency in much the same fashion as in the case of the half-wave antenna already described. However, the reactance varies at a more rapid *rate* as the applied frequency is varied; on a harmonic antenna a given percentage change in applied frequency causes a greater change in the phase of the reflected current as related to the outgoing current than is the case with a half-wave antenna. This is because, in traveling the greater length of wire in a harmonic antenna, the reflected current gains the same amount of time in *each* half-wave section, if the antenna is too short for resonance, and these gains add up as the current travels back to the driving point. When the antenna is too long, the reverse occurs and the reflected current progressively drops behind in phase as it travels back to the point at which the voltage is applied. This effect increases with the length of the antenna, and the change of phase can be quite rapid when the frequency applied to an antenna operated on a high-order harmonic is varied.

Another way of looking at it is this: Consider the antenna of Fig. 2-9A, driven at the end by a source of power having a frequency f. When f is equal to the fundamental resonant frequency of the antenna there is one half wavelength on the wire, with the current distribution as shown. At this frequency the antenna is resonant and it appears as a pure resistance of high value (because the current is small and the voltage is high at the feed point) to the source of power.

If the frequency is now increased slightly the antenna will be too long and the resultant current at the input terminals will lag behind the applied voltage (as explained by Fig. 2-6) and the antenna will have inductive reactance along with resistance. As we continue to raise the frequency the value of reactance increases to a maximum and then decreases, reaching zero when the frequency is such that the wire is an odd number of *quarter* wavelengths long. On further increasing the frequency of the reactance becomes capacitive, increases to a maximum and then decreases to zero again. At this second zero-reactance point there are two complete standing waves of current (two half waves or one wavelength) and the wire is exactly resonant on its second harmonic. This last condition is shown in Fig. 2-9B; the frequency is now $2f$, twice the original figure. In varying the frequency from f to $2f$ the resistance seen by the source of power also varies, decreasing as the frequency is raised above f and reaching a minimum when the wire is an odd number of quarter wavelengths long, then rising again with increasing frequency until it reaches a new maximum when the frequency is $2f$.

This behavior of reactance and resistance is shown in Fig. 2-10. A similar change in reactance and resistance occurs when the frequency is moved from *any* harmonic to the *next adjacent* one, as well as between the fundamental and second harmonic shown in the drawing. That is, the impedance goes through a cycle, starting with a high value of pure resistance, then becoming inductive and decreasing, passing through a low value of pure resistance, and then becoming capacitive and increasing until it again reaches a high value of pure resistance. This cycle occurs as the frequency is continu-

ously varied from any harmonic to the next higher one.

Looking now at Fig. 2-9C, the frequency has been increased to $10f$, ten times its original value, so the antenna is operated on its tenth harmonic. Raising the frequency to $11f$, the eleventh harmonic, causes the impedance of the antenna to go through the complete cycle described above. But $11f$ is only 10% higher than 10f, so a 10% change in frequency has caused a complete impedance cycle. In contrast, changing from f to $2f$ is a 100% increase in frequency—for the same impedance cycle. The impedance changes ten times as fast when the frequency is varied about the 10th harmonic as it does when the frequency is varied the same percentage about the fundamental.

To offset this, the actual impedance change—that is, the ratio of the maximum to the minimum impedance through the impedance cycle—is not as great at the higher harmonics as it is near the fundamental. This is because the radiation resistance increases with the order of the harmonic, raising the minimum point on the resistance curve and also lowering the maximum point. This comes about because the reflected current returning to the input end of a long harmonic wire is not as great as the outgoing current since energy has been lost by radiation; this is not taken into account in the theoretical pictures of current distribution so far discussed.

The over-all result, then, is that the *magnitude* of the impedance variations becomes less as the wire is operated at increasingly higher harmonics. Nevertheless, the impedance reaches a maximum at each adjacent harmonic and a minimum halfway between, independently of the actual values of impedance.

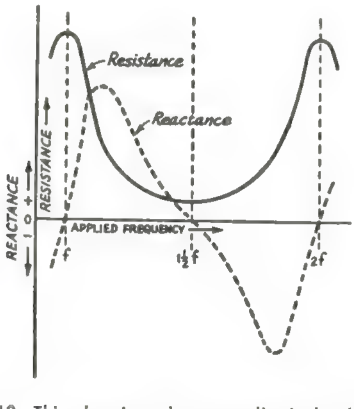


Fig. 2-10—This drawing shows qualitatively the way in which the reactance and resistance of an end-fed antenna vary as the frequency is increased from the fundamental (f) to the second harmonic (2f). Actual values of resistance and reactance and the frequencies at which the reactances are maximum will depend on the size of the conductor and the height of the antenna above ground.

OTHER ANTENNA PROPERTIES

Polarization

The polarization of a half-wave dipole is the same as the direction of its axis. That is, at distances far enough from the antenna for the waves to be considered as plane waves (see Chapter One) the direction of the electric component of the field is the same as the direction of

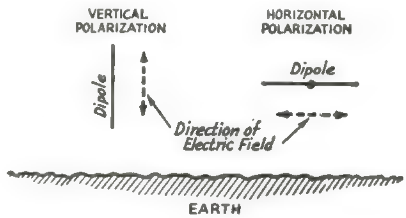


Fig. 2-11—Vertical and horizontal polarization of a dipole. The direction of polarization is the direction of the electric field with respect to earth.

the antenna wire. Vertical and horizontal polarization, the two most commonly used for antennas, are indicated in Fig. 2-11.

Antennas composed of a number of half-wave elements with all arranged so that their axes lie in the same or parallel directions will have the same polarization as that of any one of the elements. A system composed of a group of horizontal dipoles, for example, will be horizontally polarized. If both horizontal and vertical elements are used, the polarization will be the *resultant* of the contributions made by each set of elements to the total electromagnetic field at a distance. In such a case the resultant polarization will be tilted between horizontal and vertical.

Harmonic antennas also are polarized in the direction of the wire axis. However, in some combinations of harmonic wires such as the "V" and rhombic antennas described in a later chapter the polarization becomes elliptical in most directions with respect to the antenna.

As pointed out in Chapter One, sky-wave transmission usually changes the polarization of the traveling waves. The polarization of receiving and transmitting antennas in the 3-to-30 Mc. range, where almost all communication is by means of the sky wave (except for distances of a few miles), therefore need not be the same at both ends of a communication circuit. In this range the choice of polarization for the antenna is usually determined by factors such as the height of available antenna supports, the polarization of man-made r.f. noise from near-by sources, probable energy losses in near-by houses, wiring, etc., and the likelihood of interfering with neighborhood broadcast reception. Generally speaking, the most favorable solution

to these problems is achieved when the antenna is horizontally polarized.

Reciprocity in Receiving and Transmitting

The basic conditions existing when an antenna is used for radiating power are not the same as when it is used for receiving a distant signal. In the former case the electromagnetic field originates with the antenna and the waves are not plane-polarized in the immediate vicinity. In the receiving case the antenna is always far enough away from the transmitter so that the waves that the antenna intercepts are plane-polarized. This causes the current distribution in a receiving antenna to be different than in a transmitting antenna except in a few special cases. These, however, are the cases of most interest in amateur practice, since they occur when the antenna is resonant and is delivering power to a receiver.

For all practical purposes, then, the properties of a resonant antenna used for reception are the same as its properties in transmission. It has the same directive pattern in both cases, and so will deliver maximum signal to the receiver when the signal comes from the direction in which the antenna transmits best. The impedance of the antenna is the same, at the same point of measurement, in receiving as in transmitting.

In the receiving case, the antenna is to be considered as the *source* of power delivered to the receiver, rather than as the *load* for a source of power as in transmitting. Maximum output from the receiving antenna is secured when the load to which the antenna is connected is matched to the impedance of the antenna. Under these conditions half of the total power picked up by the antenna from the passing waves is delivered to the receiver and half is reradiated into space.

"Impedance matching" in the case of a receiving antenna does not have quite the same meaning as in the transmitting case; this is considered in later chapters.

The power gain (defined later in this chapter) in receiving is the same as the gain in transmitting, assuming that certain conditions are met. One such condition is that both the antenna under test and the comparison antenna (usually a half-wave antenna) work into load impedances matched to their own impedances so that maximum power is delivered in both cases. In addition, the comparison antenna should be oriented so that it gives maximum response to the signal used in the test; that is, it should have the same polarization as the incoming signal and should be placed so that its direction of maximum gain is toward the signal source.

In long-distance transmission and reception via the ionosphere the relationship between receiving and transmitting may not be exactly reciprocal. This is because the waves do not take exactly the same paths at all times and so may show considerable variation in alternate transmission and reception. Also, when more than one layer is involved in the wave travel it is sometimes possible for transmission to be good in one direction and reception to be poor in the other, over the same path. In addition, the polarization of the waves is shifted in the ionosphere, as pointed out in Chapter One. Since the tendency is for the arriving wave to be more or less horizontally polarized, regardless of the polarization of the transmitting antenna, a vertically-polarized antenna can be expected to show more difference between transmission and reception than a horizontal antenna. On the average, however, an antenna that transmits well in a certain direction will also give favorable reception from the same direction, despite ionosphere variations.

Pick-Up Efficiency

Although the transmitting and receiving properties of an antenna are, in general, reciprocal, there is another fundamental difference between the two cases that is of very great practical importance. In the transmitting case all the power supplied to the antenna is radiated (assuming negligible ohmic resistance) regardless of the physical size of the antenna system. For example, a 300-Mc. half-wave radiator, which is only about 19 *inches* long, radiates every bit as efficiently as a 3.5-Mc. half-wave antenna, which is about 134 *feet* long. But in receiving, the 300-Mc. antenna does not abstract anything like the amount of energy from passing waves that the 3.5-Mc. antenna does.

This is because the section of wave front from which the antenna can draw energy extends only about a quarter wavelength from the conductor. At 3.5 Mc. this represents an area roughly ½ wavelength or 134 feet in diameter, but at 300 Mc. the diameter of the area is only about 3 feet. Since the energy is evenly distributed throughout the wave front regardless of the wavelength, the effective area that the receiving antenna can utilize varies directly with the *square* of the wavelength. A 3.5-Mc. half-wave antenna therefore picks up something like 2500 times as much energy as a 300-Mc. half-wave antenna, the field strength being the same in both cases.

The higher the frequency, consequently, the less energy a receiving antenna has to work with. This, it should be noted, does not affect the *gain* of the antenna. In making gain measurements, both the antenna under test and the comparison antenna are working at the same frequency. Both therefore are under the same handicap with respect to the amount of energy that can be intercepted. Thus the effective area of an antenna at a given wavelength is directly proportional to its gain. Although the pick-up efficiency decreases rapidly with increasing frequency, the smaller dimensions of antenna systems in the v.h.f. and u.h.f. regions make it relatively easy to obtain high gain. This helps to overcome the loss of received signal energy.

The Induction Field

Throughout this chapter the fields we have been discussing are those forming the traveling electromagnetic waves—the waves that go long distances from the antenna. These are the radiation fields. They are distinguished by the fact that their intensity is inversely proportional to the distance and that the electric and magnetic components, although perpendicular to each other in the wave front, are in phase in time. Several wavelengths from the antenna, these are the only fields that need to be considered.

Close to the antenna, however, the situation is much more complicated. In an ordinary electric circuit containing inductance or capacitance the magnetic field is a quarter cycle out of phase (in time) with the electric field. The intensity of these fields decreases in a complex way with distance from the source. These are the induction fields. The induction field exists about an antenna along with the radiation field, but dies away with much greater rapidity as the distance from the antenna is increased. At a distance equal to the wavelength divided by 2π, or slightly less than 1/6 wavelength, the two types of field have equal intensity.

Although the induction field is of no importance insofar as effects at a distance are concerned, it *is* important when antenna elements are coupled together, particularly when the spacing between elements is small. Also, its existence must be kept in mind in making field-strength measurements about an antenna. Error may occur if the measuring equipment is set up too close to the antenna system.

Radiation Patterns And Directivity

A graph showing the actual or relative intensity, at a fixed distance, as a function of the direction from the antenna system is called a radiation pattern. At the outset it must be realized that such a pattern is a three-dimensional affair and therefore cannot be represented in a plane drawing. The "solid" radiation pattern of an antenna in free space would be found by measuring the field strength at every point on the surface of an imaginary sphere having the antenna at its center. The information so obtained is then used to construct a solid figure such that the distance from a fixed point (representing the antenna) to the surface, in any direction, is proportional to the field strength from the antenna in that direction.

THE ISOTROPIC RADIATOR

The radiation from a practical antenna never has the same intensity in all directions. The intensity may even be zero in some directions from the antenna; in others it may be greater than one would expect from an antenna that *did* radiate equally well in all directions. But even though no actual antenna radiates with equal intensity in all directions, it is nevertheless useful to assume that such an antenna exists. It can be used as a "measuring stick" for comparing the properties of actual antenna systems. Such a hypothetical antenna is called an isotropic radiator.

The solid pattern of an isotropic radiator, therefore, would be a sphere, since the field strength is the same in all directions. In any plane containing the isotropic antenna (which may be considered to be a point in space, or a "point source") the pattern is a circle with the antenna at its center. The isotropic antenna has the simplest possible directive pattern; that is, it has no directivity at all.

An infinite variety of pattern shapes, some quite complicated, is possible with actual antenna systems.

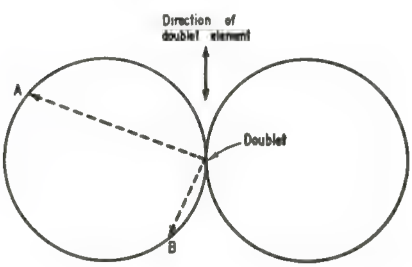


Fig. 2-12—Directive diagram of an elementary doublet in the plane containing the wire axis. The same diagram applies to any antenna considerably less than a half wavelength long. The length of each arrow represents the relative field strength in that direction.

RADIATION FROM DIPOLES

In the analysis of antenna systems it is convenient to make use of another fictitious type of antenna called an **elementary doublet** or **elementary dipole.** This is just a very short length of conductor, so short that it can be assumed that the current is the same throughout its length. (In an actual antenna, it will be remembered, the current is different all along its length.) The radiation intensity from an elementary doublet is greatest at right angles to the line of the conductor, and decreases as the direction becomes more nearly in line with the conductor until, right off the ends, the intensity is zero. The directive pattern in a single plane, one containing the conductor, is shown in Fig. 2-12. It consists of two tangent circles. The *solid* pattern is the doughnut-shaped figure described when the plane shown in the drawing is rotated about the conductor as an axis.

The radiation from an elementary doublet is not uniform in all directions because there is a definite direction to the current flow along the conductor. It will be recalled that a similar con-

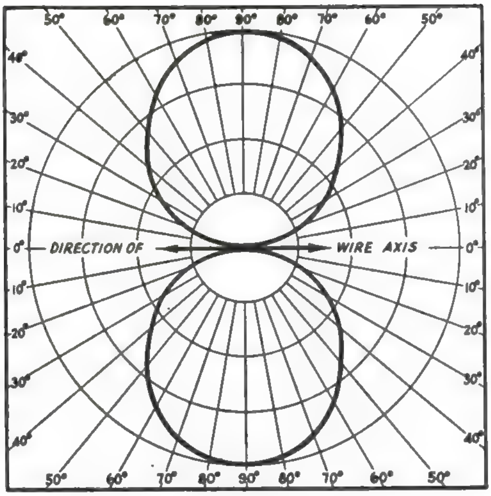


Fig. 2-13—Plane directive diagram (*E* plane) of a half-wave antenna. The solid line shows the direction of the wire, although the antenna itself is considered to be merely a point at the center of the diagram. As explained in the text, a diagram such as this is simply a cross section of the solid figure that describes the relative radiation in all possible directions. The radial scale is purely arbitrary and is proportional to the field strength (voltage). This is also true of the diagrams in Figs. 2-16, 2-17 and 2-18.

dition exists in the ordinary electric and magnetic fields set up when current flows along any conductor; the field strength near a coil, for example, is greatest at the ends and least on the outside of the coil near the middle of its length. There is nothing strange, therefore, in the idea that the field strength should depend on the direction in which it is measured from the radiator.

When the antenna has appreciable length, so that the current in every part is not the same at any given instant, the shape of the radiation pattern changes. In this case the pattern is the summation of the fields from *each* elementary doublet of which the antenna may be assumed to consist, strung together in chain fashion. If the antenna is short compared with a half wavelength there is very little change in the pattern, but at a half wavelength the pattern takes the shape shown in cross section in Fig. 2-13. The intensity decreases somewhat more rapidly, as the angle with the wire is made smaller, than in the case of the elementary doublet. This tendency continues as the wire length is increased, the shape of one half of the diagram becoming more and more compressed as the antenna length approaches a full wavelength. (The antenna is assumed to be driven at the center, as in Fig. 2-8B and 2-8C.)

The solid pattern from a half-wave wire is formed, just as in the case of the doublet, by rotating the plane diagram shown in Fig. 2-13 about the wire as an axis. If there is trouble in visualizing the solid pattern, it will help if the diagrams shown in Figs. 2-12 and 2-13 are copied on cardboard, cut out to shape, and then fastened at the center to a piece of stiff wire in the direction indicated for the antennas in the drawings. If the wire is rotated rapidly in the fingers the cardboard will appear to form a solid figure that has the same shape as the solid radiation pattern.

The directive diagrams just discussed are actually cross sections of the solid pattern, cut by planes in which the axis of the antenna lies. If the solid pattern is cut by any other plane the diagram will be different. For instance, imagine a plane passing through the center of the wire at right angles to the axis. The cross section of the pattern for either the elementary doublet or the half-wave antenna will simply be a circle in that case. This is shown in Fig. 2-14 where the dot at the center represents the antenna as viewed "end-on"; in other words, the antenna is perpendicular to the page. This means that in any direction in a plane at right angles to the wire, the field intensity is exactly the same at the same distance from the antenna. At right angles to the wire, then, an antenna a half wave or less in length is *nondirectional*. Also, at every point on such a circle the field is in the *same phase*.

E- AND H-PLANE PATTERNS

Although the solid pattern of an antenna cannot be shown adequately on a flat sheet of paper, cross-sectional or **plane diagrams** are very useful. Two such diagrams, one in the plane containing the axis of the antenna and one in the plane perpendicular to the axis, can give a great deal of information. The former is called the E-plane pattern and the latter the H-plane pattern. These designations are used because they represent the planes in which the electric (symbol E), and the magnetic (symbol H) lines of force lie, respectively. The E lines are taken to represent the polarization of the antenna, consistent with the description of antenna polarization given earlier. The electromagnetic field pictured in Fig. 1-1, as an example, is the field that would be radiated from a vertically-

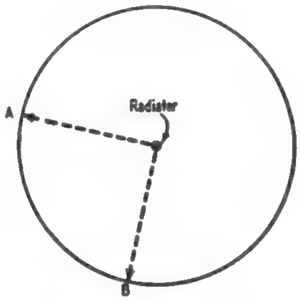


Fig. 2-14—Directive diagram of a doublet or dipole in the plane perpendicular to the wire axis (*H* plane). The direction of the wire is into or out of the page.

polarized antenna; that is, an antenna in which the conductor is mounted vertically over the earth.

After a little practice, and with the exercise of a little imagination, the complete solid pattern can be visualized with fair accuracy from inspection of just the two diagrams. Plane diagrams are plotted on polar-coordinate paper—graph paper with radial lines marking the 360 degrees of a circle, and having a linear scale along the radii for marking amplitudes.

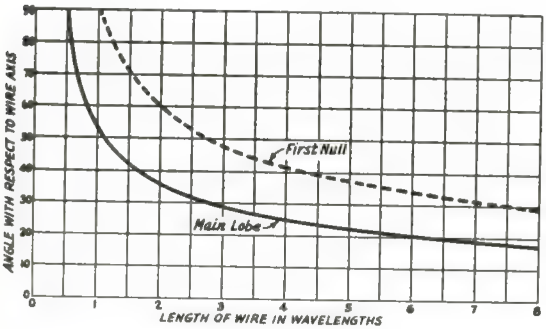


Fig. 2-15—Angle at which the field intensity from the main lobe of a harmonic antenna is maximum, as a function of the wire length in wavelengths. The curve labeled "First Null" locates the angle at which the intensity of the main lobe decreases to zero. The null marking the other boundary of the main lobe always is at zero degrees with the wire axis.

The points on the pattern where the radiation is zero are called nulls, and the curved section from one null to the next on the plane diagram, or the corresponding section on the solid pattern, is called a lobe or ear.

HARMONIC-ANTENNA PATTERNS

In view of the change in radiation pattern as the length of the antenna is increased from the elementary doublet to the half-wave dipole, it is to be expected that further pattern changes will occur as the antenna is made still longer. We find, as a matter of fact, that the patterns of harmonic antennas differ very considerably from the pattern of the half-wave dipole.

As explained earlier in this chapter, a harmonic antenna consists of a series of half-wave sections with the currents in adjacent sections always flowing in opposite directions. This type of current flow causes the pattern to be split up into a number of lobes. If there is an *even* number of half waves in the harmonic antenna there is always a null in the plane at right angles to the wire; this is because the radiation from one half wave section cancels the radiation from the next, in that particular direction. If there is an *odd* number of half waves in the antenna, the radiation from all but one of the sections cancels itself out in the plane perpendicular to the wire .The "left-over" section radiates like a half-wave dipole, and so a harmonic antenna with an odd number of half wavelengths does have some radiation at right angles to its axis.

The greater the number of half waves in a harmonic antenna, the larger the number of lobes into which the pattern splits. A feature of all such patterns is the fact that the "main" lobe—the one that gives the largest field strength at a given distance—always is the one that makes the smallest angle with the antenna wire. Furthermore, this angle becomes smaller as the length of the antenna is increased. Fig. 2-15 shows how the angle which the main lobe makes with the axis of the antenna varies with the antenna length in wavelengths. The angle shown by the solid curve is the maximum point of the lobe; that is, the direction in which the field strength is greatest. The broken curve shows the angle at which the first null (the one that occurs at the smallest angle with the wire) appears. There is also a null in the direction of the wire itself (0 degrees) and so the total width of the main lobe is the angle between the wire and the first null. It can be seen from Fig. 2-15 that the width of the lobe decreases as the wire becomes longer. At 1 wavelength, for example, it has a width of 90 degrees, but at 8 wavelengths the width is slightly less than 30 degrees.

Plane diagrams of the radiation patterns of several harmonic wires are shown in Figs. 2-16 to 2-18, inclusive. These are free-space diagrams in the plane containing the wire axis (*E* plane), corresponding to the diagrams for the elementary doublet and half-wave dipole shown in Figs. 2-12 and 2-13. They are based on an infinitely-thin antenna conductor with ideal current distribution, and in a practical antenna sys-

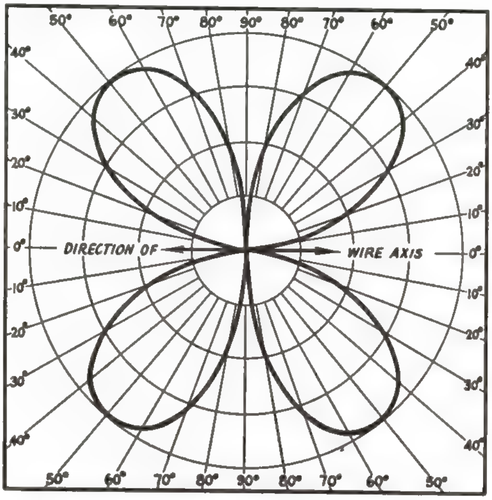


Fig. 2-16—Free-space directive diagram of a 1-wavelength harmonic antenna in the plane containing the wire axis.

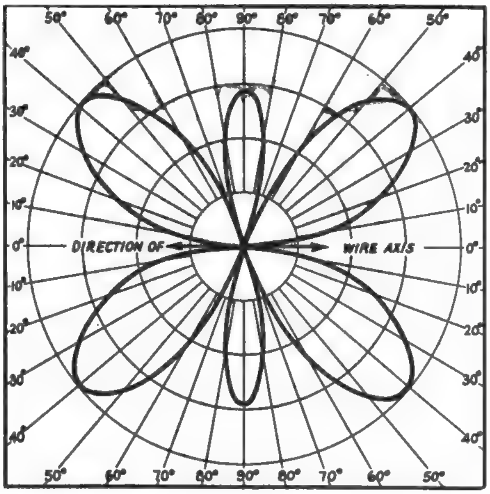


Fig. 2-17—Free-space directive diagram of a 1½-wavelength harmonic antenna in the plane containing the wire axis.

tem will be modified by the presence of the earth and other effects that will be considered later. However, these few diagrams do show the tendency for the pattern to break up into more and more lobes as the antenna is made longer, and also show that these "secondary" lobes have progressively smaller amplitude as compared with the main lobe.

HOW PATTERNS ARE FORMED

The radiation pattern of the half-wave dipole is found by summing up, at every point on the surface of a sphere with the antenna at its center, the field contributions of all the elementary dipoles that can be imagined to make up the full-size dipole. Antenna systems often are composed of a group of half-wave dipoles arranged in various ways, in which case each half-wave dipole is called an antenna element. An antenna having two or more such dipoles is called a multielement antenna. (A harmonic antenna can be considered to be constructed of a number of such elements connected in series and fed power appropriately, as described earlier, but is not usually classed as a multielement antenna.)

In a multielement antenna system the over-all radiation pattern is determined by the way in which the fields at a distant point from the separate antenna elements combine. With two antenna elements, for example, the field strength at a given point depends on the amplitudes and phase relationship of the fields from each antenna. A requirement in working out a radiation pattern is that the field strength be measured or calculated at a *distant* point—distant enough so that, if the elements carry equal currents, the field strength from each is exactly the same even though the size of the antenna system may be such that one antenna element is a little nearer the measuring point than another. On the other hand, this slight difference in distance, even though it may be only a small fraction of a wave-length in many miles, is very important in determining the *phase* relationships between the fields from the various elements.

The principle on which the radiated fields combine to produce the directive pattern, in the case of multielement antennas, is illustrated in the simple example shown in Fig. 2-19. In this case it is assumed that there are two antenna elements, each having a circular directive pattern. The two elements therefore could be half-wave dipoles oriented perpendicular to the page (which gives the plane pattern shown in Fig. 2-14). The separation between the two elements is assumed to be a half-wavelength, and the currents in them are assumed to be equal. Furthermore, the two currents are in phase; that is, they reach their maximum values in the same polarity at the same instant.

Under these conditions the fields from the two antennas will be in the same phase at any point that is equally distant from both antenna elements. At the instant of time selected for the drawing of Fig. 2-19 the solid circles having the upper antenna at their centers represent, let us say, the location of all points at which the field intensity is maximum and has the direction indicated by the arrowheads. The *distance* between each pair of concentric solid circles, measured along a radius, is equal to one wavelength because, as described earlier in this chapter, it is only at intervals of this distance that the fields are in phase. The broken circle locates the points at which the field intensity is the same as in the case of the solid circles, but is *oppositely* directed. It is therefore 180 degrees out of phase with the field denoted by the solid circles, and the distance between the solid and broken circles is therefore one-half wavelength.

Similarly, the solid circles centered on the

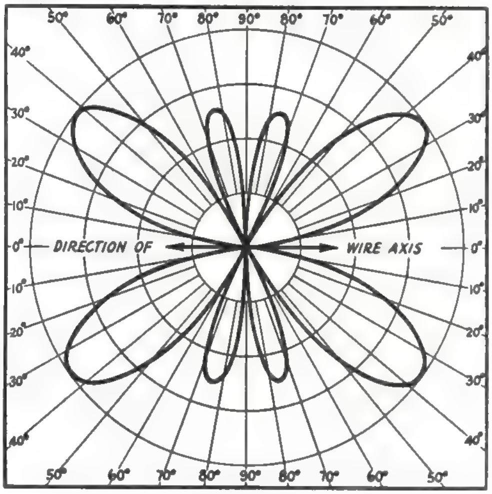


Fig. 2-18—Free-space directive diagram of a 2-wavelength harmonic antenna in the plane containing the wire axis.

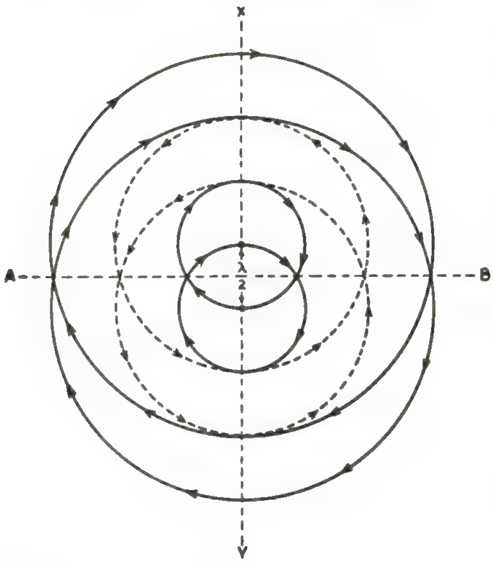


Fig. 2-19—Interference between waves from two separated radiators causes the resultant directional effects to differ from those of either radiator alone. The two radiators shown here are separated one-half wavelength. The radiation fields of the two cancel along the line XY but, at distances which are large compared with the separation between the radiators, add together along line AB. The resultant field decreases uniformly as the line is swung through intermediate positions from AB to XY.

lower antenna locate all points at which the field intensity from that antenna is maximum and has the same direction as the solid circles about the upper antenna. In other words, these circles represent points in the same phase as the solid circles around the upper antenna. The broken circle having the lower antenna at its center likewise locates the points of opposite phase.

Considering now the fields from both antennas, it can be seen that along the line *AB* the fields from the two always are exactly in phase, because every point along *AB* is equally distant from both antenna elements. However, along the line *XY* the field from one antenna always is *out* of phase with the other, because every point along *XY* is a half-wavelength nearer one element than the other. It takes one-half cycle longer, therefore, for the field from the more distant element to reach the same point as the field from the nearer antenna, and thus the one field arrives 180 degrees out of phase with the other. Since we have assumed that the points considered are sufficiently distant so that the amplitudes of the fields from the two antennas are the same, the *resultant* field at any point along *XY* is zero and the antenna combination shown will have a null in that direction. However, the two fields add together along *AB*, and the field strength in that direction will be twice the amplitude of the field from either antenna alone.

The drawing of Fig. 2-19 is not quite accurate because it cannot be made large enough. Actually, the two fields along *AB* do not have exactly the same direction until the distance to the measuring point is large enough, compared with the dimensions of the antenna system, so that the waves become plane. In a drawing of limited size the waves are necessarily represented as circles—that is, as representations of a *spherical* wave. The reader, therefore, should imagine Fig. 2-19 as being so much enlarged that the circles crossing *AB* are substantially straight lines in the region under discussion.

Pattern Construction

The drawing of Fig. 2-19 does not tell us much about what happens to the field strength at points that do not lie on either *AB* or *XY*, although we could make the reasonable guess that the field strength at intermediate points probably would decrease as the point was moved along the arc of a circle farther away from *AB* and nearer to *XY*. To construct an actual pattern it is necessary to use a different method. It is simple in principle and can be done with a ruler, protractor and pencil.

In Fig. 2-20 the two antennas, *A* and *B*, are assumed to have circular radiation patterns, and to carry equal currents in the same phase. (In other words, the conditions are the same as in Fig. 2-19.) The relative field strength at a distant point *P* is to be determined. Here again the limitations of the printed page make it necessary to use the imagination, because we assume that P is far enough from *A* and *B* so that the lines *AP* and *BP* are, for all practical purposes, parallel. When this is so, the distance *d*, between *B*

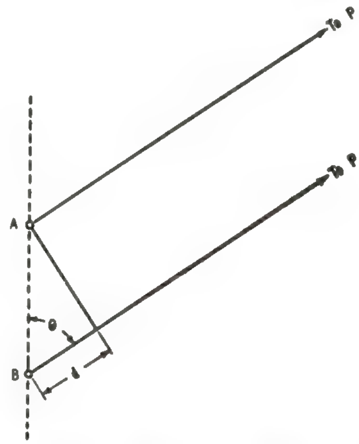


Fig. 2-20—Graphical construction to determine the relative phase, at a distant point, of waves originating at two antennas, *A* and *B*. The phase is determined by the additional distance, *d*, that the wave from *B* has to travel to reach the distant point. This distance will vary with the angle that the direction to *P* makes with the axis of the antenna system.

and a perpendicular dropped to BP from A, will be equal to the difference in length between the distance from A to P and the distance from B to P. The distance d thus measures the difference in the distances the waves from A and B have to cover to reach P; d is also, therefore, a measure of the difference in the time of arrival or *phase* of the waves at P.

Under the assumed conditions, the relative field strengths easily can be combined graphi-

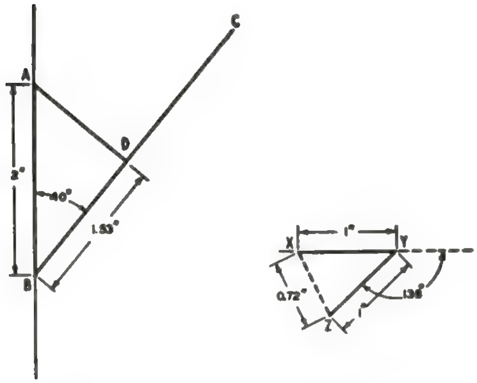


Fig. 2-21—Graphical construction in the example discussed in the text.

cally. The phase angle in degrees between the two fields at P is equal to

$$\frac{d}{\lambda} \times 360$$

where λ is the wavelength and d is found by constructing a figure similar to that shown in Fig. 2-20 for P in any desired direction. The angle θ is the angle between a line to P and the line drawn between the two antenna elements, and is used simply to identify the direction of P from the antenna system. λ and d must be expressed in the same length units.

For example, let us assume that θ is 40 degrees. We then arbitrarily choose a scale such that four inches is equal to one wavelength—a scale large enough for reasonable accuracy but not too large to be unwieldy. Since the two antenna elements are assumed to be a half-wavelength apart, we start the drawing by placing two points two inches apart and connecting them by a line, as shown in Fig. 2-21. Then, using B as a center and employing the protractor, we lay off an angle of 40 degrees and draw the line BC. The next step is to drop a perpendicular from A to BC; this may be done with the 90-degree mark on the protractor but the corner of an ordinary sheet of paper will do just about as well. The distance BD is then measured, preferably with a ruler graduated in tenths of inches rather than the more usual eighths. By actual measurement distance BD is found to be 1.53 inches. The phase difference is therefore $d/\lambda \times 360 = 1.53/4 \times 360 = 138$ degrees.

The relative field strength in the direction given by θ (40 degrees in this example) is found by arbitrarily selecting a line length to represent the strength of the field from each antenna, and then combining them "vectorially." One inch is a convenient length to select. XY, Fig. 2-21, is such a line, representing the strength of the field from antenna element A. We then measure off an angle of 133 degrees from XY, using Y as a center, and draw YZ one inch long to represent the strength and phase of the field from antenna element B. The angle is measured off clockwise from XY because the field from B lags behind that from A. The distance from X to Z then represents the relative field strength resulting from the combination of the separate fields from the two antennas, and measurement shows it to be approximately 0.72 inch. In the direction θ, therefore, the field strength is 72% as great as the field from either antenna alone.

By selecting different values for θ and proceeding as above in each case, the complete pattern can be determined. When θ is 90 degrees, the phase difference is zero and YZ *and* XY are simply end-to-end along the same line. The maximum field strength is therefore twice that of either antenna alone. When θ is zero, YZ lies on top of XY (phase difference 180 degrees) and the distance XZ is therefore zero; in other words, the radiation from B cancels that from A at such an angle.

The patterns of more complex antenna systems can readily be worked out by this method, although more labor is required if the number of elements is increased. But whether or not actual patterns are worked out, an understanding of the method will do much to make it plain why certain combinations of antenna elements result in specific directive patterns.

The illustration above is a very simple case, but it is only a short step to systems in which the antenna elements do not carry equal currents or currents in the same phase. A difference in current amplitude is easily handled by making the lengths of lines XY and YZ proportional to the current in the respective elements; if the current in B is one-half that in A, for example, YZ would be drawn one-half as long as XY. If B's current leads the current in A by 25 degrees, then after the angle determined by the distance d is found the line YZ is simply rotated 25 degrees in the *counterclockwise* direction before measuring the distance XZ. The rotation would be *clockwise* for any line representing a lagging current. The lead or lag of current always has to be referred to the current in *one* element of the system, but any desired element can be chosen as the reference.

DIRECTIVITY AND GAIN

It has been stated that all antennas, even the simplest types, exhibit directive effects in that the intensity of radiation is not the same in all directions from the antenna. This property of radiating more strongly in some directions than in others is called the directivity of the antenna. It can be expressed quantitatively by comparing

the solid pattern of the antenna under consideration with the solid pattern of the isotropic antenna. The field strength (and thus power per unit area or "power density") will be everywhere the same on the surface of an imaginary sphere having a radius of many wavelengths and having an isotropic antenna at its center. At the surface of the same imaginary sphere around an actual antenna radiating the same total power, the directive pattern will result in greater power density at some points and less at others. The ratio of the maximum power density to the average power density taken over the entire sphere (the latter is the same as from the isotropic antenna under the specified conditions) is the numerical measure of the directivity of the antenna. That is,

$$D = \frac{P}{P_{av.}}$$

where D is the directivity, P is the power density at its maximum point on the surface of the sphere, and $P_{av.}$ is the average power density.

Gain

The gain of an antenna is closely related to its directivity. Since directivity is based solely on the *shape* of the directive pattern, it does not take into account any power losses that may occur in an actual antenna system. In order to determine gain these losses must be subtracted from the power supplied to the antenna. The loss is normally a constant percentage of the power input, so the antenna gain is

$$G = k\frac{P}{P_{av.}}$$

where G is the gain expressed as a power ratio, k is the efficiency (power radiated divided by power input) of the antenna, and P and P_{av} are as above. For many of the antenna systems used by amateurs the efficiency is quite high (the loss amounts only to a few percent of the total), and in such cases the gain is essentially equal to the directivity.

The more the directive diagram is compressed —or, in common terminology, the "sharper" the lobes—the greater the power gain of the antenna. This is a natural consequence of the fact that as power is taken away from a larger and larger portion of the sphere surrounding the radiator it is added to the smaller and smaller volume represented by the lobes. The power is therefore concentrated in some directions at the expense of others. In a general way, the smaller the volume of the solid radiation pattern, compared with the volume of a sphere having the same radius as the length of the largest lobe in the actual pattern, the greater the power gain.

Gain referred to an isotropic radiator is necessarily theoretical; that is, it has to be calculated rather than measured because the isotropic radiator has no existence. In practice, measurements on the antenna being tested usually are compared with measurements made on a half-wave dipole. The latter should be at the same height and have the same polarization as the antenna under test, and the reference field—that from the half-wave dipole comparison antenna—should be measured in the most favored direction of the dipole. The data can be secured either by measuring the field strengths produced at the same distance from both antennas when the same power is supplied to each, or by measuring the power required in each antenna to produce the same field strength at the same distance.

A half-wave dipole has a theoretical gain of 2.14 db. over an isotropic radiator. Thus the gain of an actual antenna over a half-wave dipole can be referred to isotropic by adding 2.14 db. to the measured gain, or if the gain is expressed over an isotropic antenna it can be referred to a half-wave dipole by subtracting 2.14 db.

It should be noted that the field strength (voltage) produced by an antenna at a given point is proportional to the square root of the power. That is, when the two are expressed as ratios (the usual case),

$$\frac{F_1}{F_2} = \sqrt{\frac{P_1}{P_2}}$$

Antenna patterns often are plotted in terms of relative field strength, and if these are to be interpreted in power the field-strength ratio

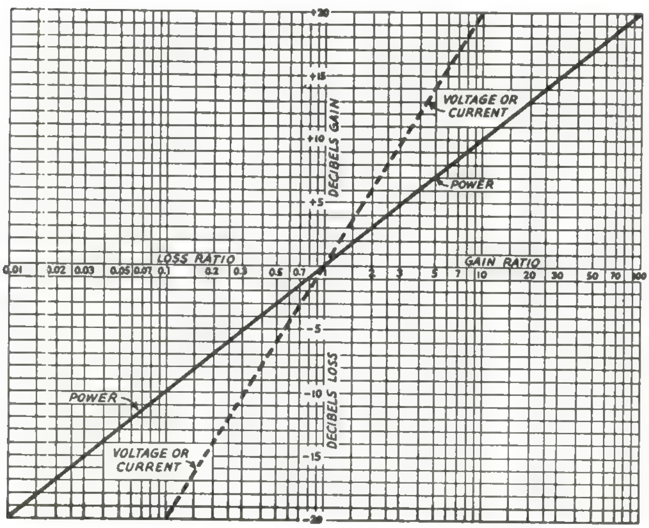


Fig. 2-22—Chart of decibels vs. power or voltage gain or loss. When the voltage curve is used the voltages must be measured across identical impedances. The range of the chart can be extended by adding (or subtracting, if a loss) 10 db. each time the power ratio is multiplied or divided by 10. or 20 db. each time the voltage is multiplied or divided by 10.

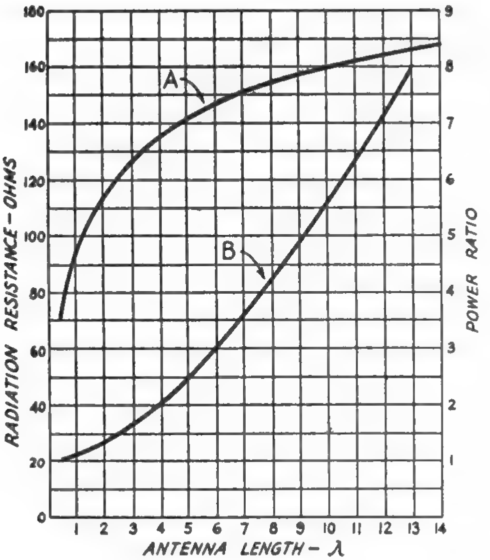


Fig. 2-23—The variation in radiation resistance and power in the major lobe of long-wire antennas. Curve A shows the change in radiation resistance with antenna length, while Curve B shows the power in the lobes of maximum radiation for long-wire antennas as a ratio to the maximum of a half-wave antenna.

must be squared. For example, if the field strength is doubled, the power ratio is 2^2, or 4.

The Decibel

As a convenience, the power gain of an antenna system is usually expressed in decibels. The decibel is an excellent practical unit for measuring power ratios because it is more closely related to the actual effect produced than the power ratio itself. One decibel represents a just-detectable *change* in signal strength, regardless of the actual value of the signal voltage. A 20-decibel (20-db.) increase in signal, for example, represents 20 observable "steps" in increased signal. The power ratio (100 to 1) corresponding to 20 db. would give an entirely exaggerated idea of the improvement in communication to be expected. The number of decibels corresponding to any power ratio is equal to 10 times the common logarithm of the power ratio, or

$$\text{db.} = 10 \log \frac{P_1}{P_2}$$

If the *voltage* ratio is given, the number of decibels is equal to 20 times the common logarithm of the ratio. That is,

$$\text{db.} = 20 \log \frac{E_1}{E_2}$$

When a voltage ratio is used both voltages must be measured in the same value of impedance. Unless this is done the decibel figure is meaningless, because it is fundamentally a measure of a *power* ratio.

The chart of Fig. 2-22 shows the number of decibels corresponding to various power and voltage ratios. One advantage of the decibel is that successive power gains expressed in decibels may simply be added together. Thus a gain of 3 db. followed by a gain of 6 db. gives a total gain of 9 db. In ordinary power ratios, the ratios would have to be multiplied together to find the total gain. Furthermore, a *reduction* in power is handled simply by subtracting the requisite number of decibels. Thus reducing the power to ½ is the same as *subtracting* 3 decibels. We might, for example, have a power gain of 4 in one part of a system and a reduction of ½ in another part, so that the total power gain is 4 × ½ = 2. In decibels, this would be 6 − 3 = 3 db. A power reduction or "loss" is simply indicated by putting a negative sign in front of the appropriate number of decibels.

Power Gains of Harmonic Antennas

In splitting off into a series of lobes, the solid radiation pattern of a harmonic antenna is compressed into a smaller volume as compared with the single-lobed pattern of the half-wave dipole. This means that there is a concentration of power in certain directions with a harmonic antenna, particularly in the main lobe. The result is that a harmonic antenna will produce an increase in field strength, in its most favored direction, over a half-wave dipole in *its* most favored direction, when both antennas are supplied with the same amount of power.

The power gain from harmonic operation is small when the antenna is small in terms of wavelengths, but is quite appreciable when the antenna is fairly long. The theoretical power gain of harmonic antennas or "long wires" is shown by curve B in Fig. 2-23, using the half-wave dipole as a base. A 1-wavelength or "second harmonic" antenna has only a slight power gain, but an antenna 9 wavelengths long will show a power increase of 5 times over the dipole. This gain is secured in one direction by eliminating the power radiated in other directions; thus the longer the wire the more directive the antenna becomes.

Curve A in Fig. 2-23 shows how the radiation resistance, as measured at a current loop, varies with the length of a harmonic antenna.

Ground Effects

REFLECTION FROM THE GROUND

The performance of an antenna, particularly with respect to its directive properties, is considerably modified by the presence of the earth underneath it. The earth acts like a huge reflector for those waves that are radiated from the antenna at angles lower than the horizon. These downcoming waves strike the surface and are reflected by a process very similar to that by which light waves are reflected from a mirror.

As in the case of light waves, the angle of reflection is the same as the angle of incidence, so that a wave striking the surface at an angle of, for instance, 15 degrees, is reflected upward from the surface at the same angle.

The reflected waves combine with the direct waves (those radiated at angles above the horizontal) in various ways, depending upon the orientation of the antenna with respect to earth, the height of the antenna, its length, and the character of the ground. At some vertical angles above the horizontal the direct and reflected waves may be exactly in phase—that is, the maximum field strengths of both waves are reached at the same time at the same spot, and the directions of the fields are the same. In such a case the resultant field strength is simply equal to the sum of the two. At other vertical angles the two waves may be completely out of phase—that is, the fields are maximum at the same instant and the directions are opposite, at the same spot. The resultant field strength in that case is the *difference* between the two. At still other angles the resultant field will have intermediate values. Thus the effect of the ground is to increase the intensity of radiation at some vertical angles and to decrease it at others.

The effect of reflection from the ground is shown graphically in Fig. 2-24. At a sufficiently large distance, two rays converging at the distant

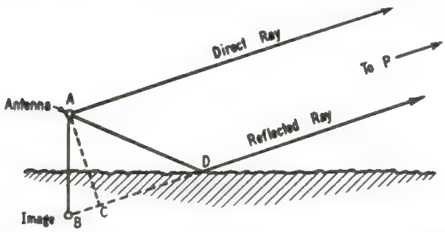


Fig. 2-24—At any distant point, P, the field strength will be the resultant of two rays, one direct from the antenna, the other reflected from the ground. The reflected ray travels farther than the direct ray by the distance BC, where the reflected ray is considered to originate at the "image" antenna.

point can be considered to be parallel. However, the reflected ray travels a greater distance in reaching *P* than the direct ray does, and this difference in path length accounts for the effect described in the preceding paragraph. If the ground were a perfect conductor for electric currents reflection would take place without a change in phase when the waves are vertically polarized. Under similar conditions there would be a complete reversal (180 degrees) of phase when a horizontally-polarized wave is reflected. The actual earth is of course not a perfect conductor, but is usually assumed to be one for purposes of calculating the vertical pattern of an antenna. The error is small except at very low vertical angles.

As an example, when the path of the reflected ray is exactly a half wave longer than the path of the direct ray, the two waves will arrive out of phase if the polarization is vertical. This corresponds to the condition illustrated in Fig. 2-19 along the line *XY*. However, if the path of the reflected ray is just a wavelength longer than that of the direct ray, the two rays arrive in phase.

Image Antennas

It is often convenient to use the concept of an image antenna to show the effect of reflection. As Fig. 2-24 shows, the reflected ray has the same path length (*AD* equals *BD*) that it would if it originated at a second antenna, of the same characteristics as the real antenna, but situated below the ground just as far as the actual antenna is above it. Like an image in a mirror, this image antenna is "in reverse," as shown in Fig. 2-25.

If the real antenna is horizontal, and is instantaneously charged so that one end is positive and the other negative, then the image antenna, also horizontal, is oppositely poled; the end under the positively-charged end of the real antenna is negative, and vice versa. Likewise, if the lower end of a half-wave vertical antenna is instantaneously positive, the end of the vertical image antenna nearest the surface is negative. Now if we look at the antenna and its image from a remote point on the surface of the ground, it will be obvious that the currents in the horizontal antenna and its image are flowing in opposite directions, or are 180 degrees *out* of phase, but the currents in the vertical antenna and its image are flowing in the *same* direction, or are *in* phase. The effect of ground reflection, or the image antenna, is therefore different for horizontal and vertical half-wave antennas. The physical reason for this difference is the fact that vertically-polarized waves are reflected from a perfectly-conducting earth with no change in phase, but that horizontally-polarized waves have their phase shifted by 180 degrees on reflection.

Reflection Factor

The effect of reflection can be expressed as a factor which, when multiplied by the free space figure for relative intensity of radiation at a given vertical angle from an antenna, gives the resultant relative radiation intensity at that same angle. The limiting conditions are those represented by the direct ray and reflected ray being exactly in phase and exactly out of phase when both, assuming there are no ground losses, have exactly equal amplitudes. Thus the resultant field strength at a distant point may be either twice the field strength from the antenna alone, or zero.

The way in which the reflection factor (based on perfectly-conducting ground) varies with antenna height is shown in the series of graphs, Figs. 2-26 to 2-43. Figs. 2-26 to 2-37 apply to horizontal antennas of any length, and to vertical antennas an *even* number of half waves long. Figs. 2-38 to 2-43 apply to vertical antennas an

odd number of half waves long. Comparing the two sets, it is seen that the positions of nulls (multiplying factor zero) and maxima (multiplying factor 2) are interchanged for the two sets of conditions.

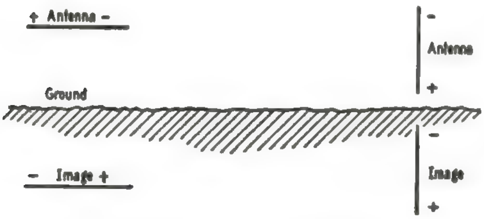


Fig. 2-25—Horizontal and vertical half-wave antennas and their images.

It must be remembered that these graphs are not plots of vertical patterns of antennas, but represent simply multiplying factors representing the result of reflection from the ground. With the distinction between vertical and horizontal antennas noted, the graphs apply equally well to *all* antennas. Also, it should be understood that they apply at *vertical* angles only. The ground, if of uniform characteristics, makes no distinction between geographical directions—that is, horizontal angles from the antenna—in reflecting waves.

Fig. 2-44 shows the angles at which nulls and maxima occur as a function of the height of the antenna. This chart gives a rough idea of the ground-reflection pattern for heights intermediate to those shown in detail in Figs. 2-26 to 2-43. It also facilitates picking the right height for any desired angle of radiation.

GROUND CHARACTERISTICS

As already explained, the charts of Figs. 2-26 to 2-43 are based on the assumption that the earth is a perfect conductor. The actual ground is far from being "perfect" as a conductor of electricity. Actually, its behavior depends considerably on the transmitted frequency. At low frequencies—through the standard broadcast

Factors by which the free-space radiation pattern of a horizontal antenna should be multiplied to include the effect of reflection from perfectly-conducting ground. These factors affect only the vertical angle of radiation (wave angle).

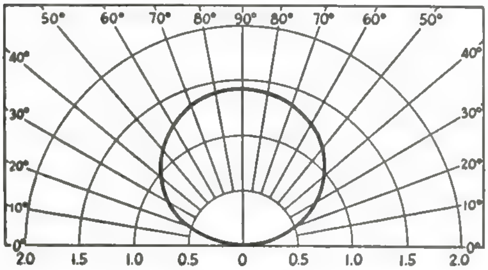


Fig. 2-26—Horizontal antennas ⅛ wavelength high.

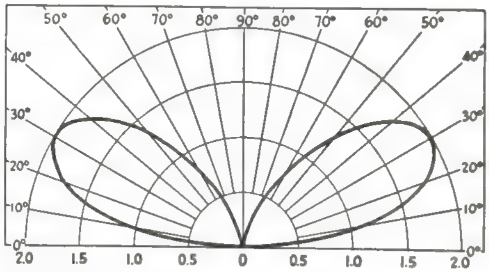


Fig. 2-29—Horizontal antennas ½ wavelength high.

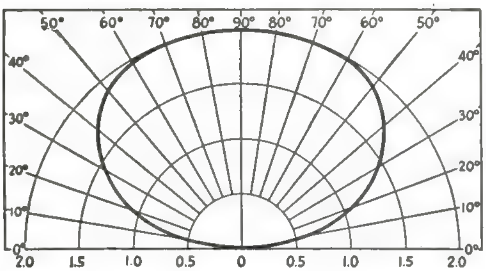


Fig. 2-27—Horizontal antennas ¼ wavelength high.

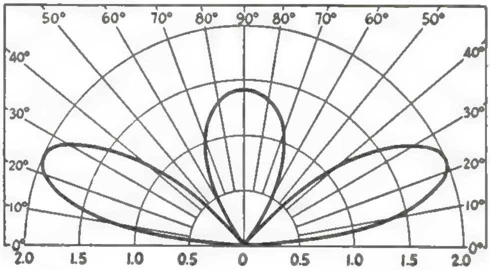


Fig. 2-30—Horizontal antennas ⅝ wavelength high.

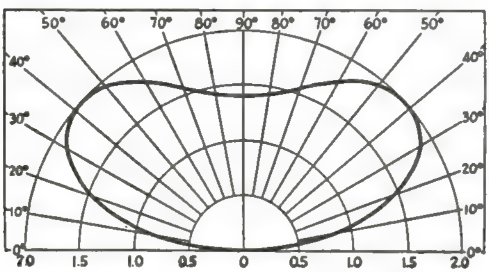


Fig. 2-28—Horizontal antennas ⅜ wavelength high.

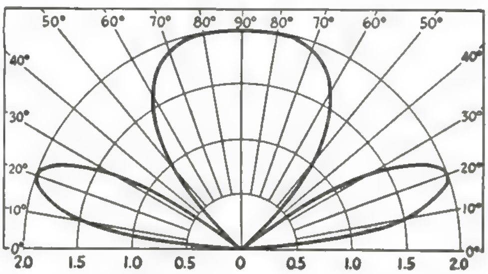


Fig. 2-31—Horizontal antennas ¾ wavelength high.

band, for example—most types of ground do act very much like a good conductor. At these frequencies the waves can penetrate for quite a distance and thus find a large cross section in which to cause current flow along their paths. The resistance of even a moderately-good conductor will be low if its cross section is large enough. The ground acts as a fairly good conductor even at frequencies as high as the 3.5-Mc. band, and so the charts give a rather good approximation of the effect of the ground at this frequency.

In the higher-frequency region the penetration decreases and the ground may even take on the characteristics of a lossy dielectric rather than a good conductor. The chief effect of this change is to absorb most of the energy radiated at the very low angles, in the frequency region from about 7 to 21 Mc. In general, the reflection factor will be lower than given by the charts at angles of less than about 10 degrees, and it is generally considered that the radiation below about 3 degrees is very small compared with the radiation at higher angles. This applies to both vertical and horizontal antennas, so that the "zero-angle" reflection factor with a vertical half-wave antenna, theoretically 2 as shown by the charts, actually is a small fraction. Thus the apparent advantage of the vertical antenna for very low-angle radiation is not realized in practice in this frequency range.

The "effective reflecting plane" of the ground —that is, the surface from which the reflection is considered to take place at the heights given in the charts—seldom coincides with the actual surface of the ground. Usually it will be found that this plane appears to be a few feet below the surface; in other words, the height of the antenna taken for purposes of estimating reflection is a few feet more than the actual height of the antenna. A great deal depends upon the character of the ground, and in some cases the reflecting plane may be "buried" a surprising distance. Thus in some instances the charts will not give an accurate indication of the effect of reflection. On the average, however, they will give a reasonably satisfactory representation of reflection effects, with the qualifications with

Factors by which the free-space radiation pattern of a horizontal antenna should be multiplied to include the effect of reflection from perfectly-conducting ground. These factors affect only the vertical angle of radiation (wave angle).

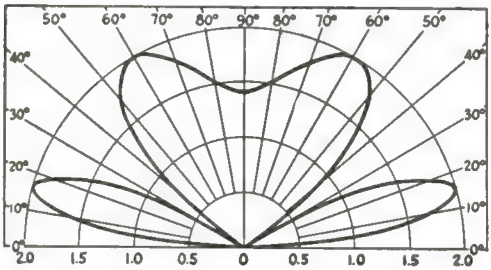


Fig. 2-32—Horizontal antennas ⅞ wavelength high.

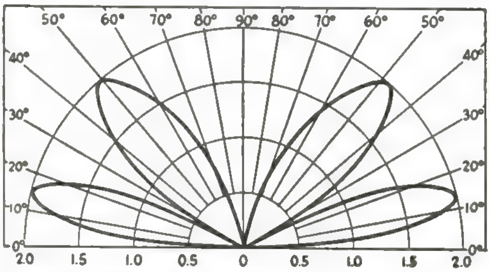


Fig. 2-23—Horizontal antennas 1 wavelength high.

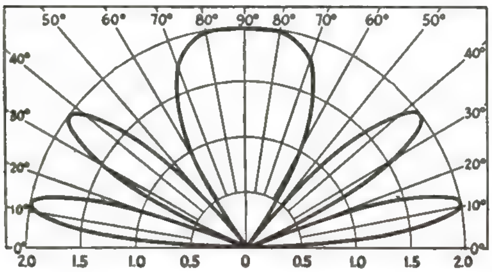


Fig. 2-34—Horizontal antennas 1¼ wavelengths high.

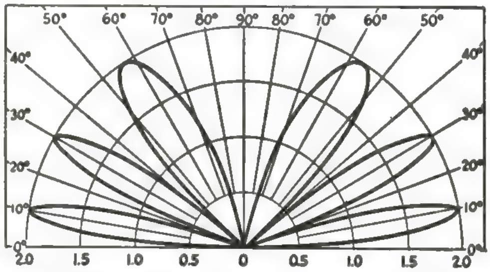


Fig. 2-35—Horizontal antennas 1½ wavelengths high.

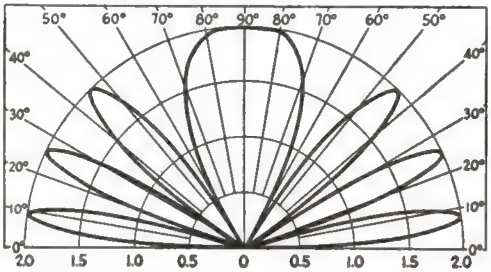


Fig. 2-36—Horizontal antennas 1¾ wavelengths high.

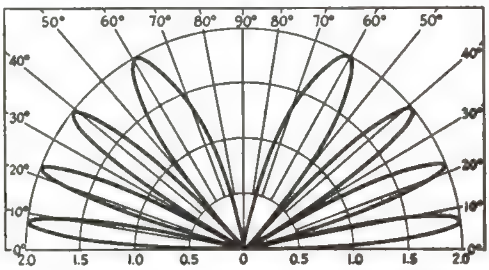


Fig. 2-37—Horizontal antennas 2 wavelengths high.

Factors by which the free-space radiation pattern of a half-wave vertical antenna should be multiplied to include the effect of reflection from perfectly-conducting ground. These factors affect only the vertical angle of radiation (wave angle).

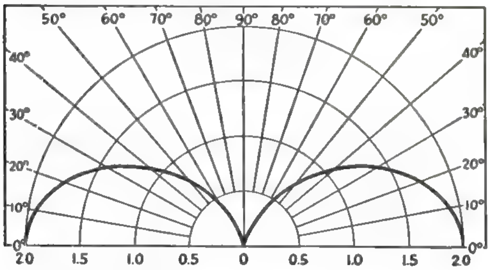


Fig. 2-38—Vertical dipole antenna with center ¼ wavelength high.

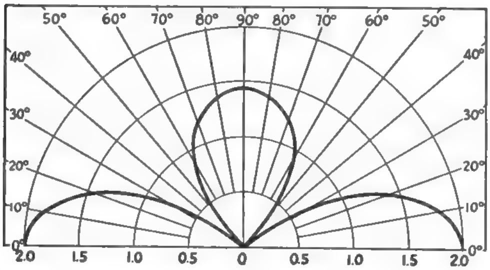


Fig. 2-39—Vertical dipole antenna with center ⅜ wavelength high.

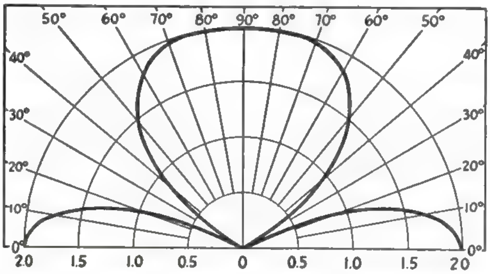


Fig. 2-40—Vertical dipole antenna with center ½ wavelength high.

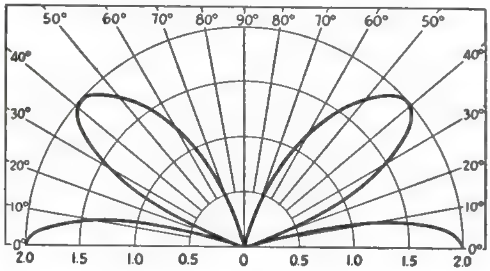


Fig. 2-41—Vertical dipole antenna with center ¾ wavelength high.

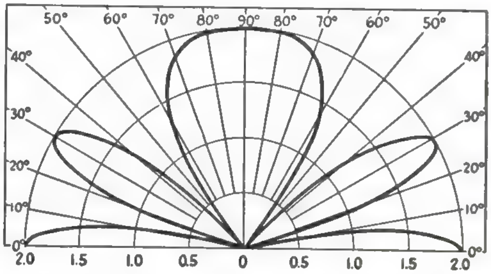


Fig. 2-42—Vertical dipole antenna with center 1 wavelength high.

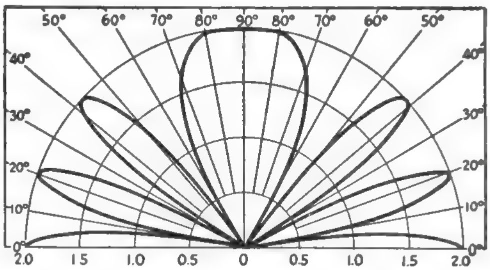


Fig. 2-43—Vertical dipole antenna with center 1½ wavelengths high.

respect to high frequencies and low angles mentioned above.

In the v.h.f. and u.h.f. region (starting in the vicinity of the 28-Mc. band) a different situation exists. At these frequencies little, if any, use is made of the part of the wave that travels in contact with the ground. The antennas, both transmitting and receiving, usually are rather high in terms of wavelength. The wave that is actually used—at least for line-of-sight communication—is in most cases several wavelengths above the surface of the ground. At such a height there is no consequential loss of energy; the direct ray travels from the transmitter to the receiver with only the normal attenuation caused by spreading, as explained in Chapter One. The loss of energy in the reflected ray is beneficial rather than otherwise, as also explained in that chapter. The net result is that radiation at very low angles is quite practicable in this frequency region. Also, there is little practical difference between horizontal and vertical polarization.

GROUND REFLECTION AND RADIATION RESISTANCE

Waves radiated from the antenna directly downward reflect vertically from the ground and, in passing the antenna on their upward journey, induce a current in it. The magnitude and phase of this induced current depends upon the height of the antenna above the reflecting surface.

The total current in the antenna thus consists of two components. The amplitude of the first is determined by the power supplied by the

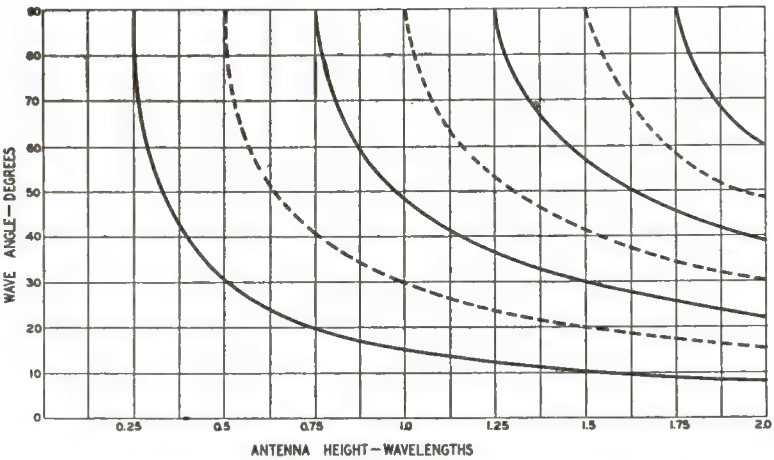


Fig. 2-44—Angles at which nulls and maxima (factor = 2) in the ground-reflection factor appear for antenna heights up to two wavelengths. The solid lines are maxima, dashed lines nulls, for all horizontal antennas and for vertical antennas having a length equal to an even multiple of one-half wavelength. For vertical antennas an odd number of half waves long, the dashed lines are maxima and the solid lines nulls. For example, if it is desired to have the ground reflection give maximum reinforcement of the direct ray from a horizontal antenna at a 20-degree wave angle (angle of radiation) the antenna height should be 0.75 wavelength. The same height will give a null at 42 degrees and a second maximum at 90 degrees.

transmitter and the free-space radiation resistance of the antenna. The second component is induced in the antenna by the wave reflected from the ground. The second component, while considerably smaller than the first at most useful antenna heights, is by no means inappreciable. At some heights the two components will be more or less in phase, so the total current is larger than would be expected from the free-space radiation resistance. At other heights the two components are out of phase, and at such heights the total current is the difference between the two components.

Thus merely changing the height of the antenna above ground will change the amount of current flow, assuming that the power input to the antenna is held constant. A higher current at the same value of power means that the effective resistance of the antenna is lower, and vice versa. In other words, the radiation resistance of the antenna is affected by the height of the antenna above ground. Fig. 2-45 shows the way in which the radiation resistance of a horizontal half-wave antenna varies with height, in terms of wavelengths, over perfectly conducting ground. Over actual ground the variations will be smaller, and tend to become negligible as the height approaches a half wavelength. The antenna on which this chart is based is assumed to have an infinitely-thin conductor, and thus has a somewhat higher free-space value of radiation resistance (73 ohms) than an antenna constructed of wire or tubing. (See Fig. 2-7.)

Ground Screens

The effect of a perfectly-conducting ground can be simulated, under the antenna, by installing a metal screen or mesh (such as chicken wire) near or on the surface of the ground. The screen preferably should extend at least a half wavelength in every direction from the antenna. Such a screen will rather effectively establish the height of the antenna insofar as radiation resistance is concerned, since it substitutes for the actual earth underneath the antenna.

For vertical quarter-wave antennas the screen also reduces losses in the ground near the antenna, since if the screen conductors are solidly bonded to each other the resistance is much lower than the resistance of the ground itself.

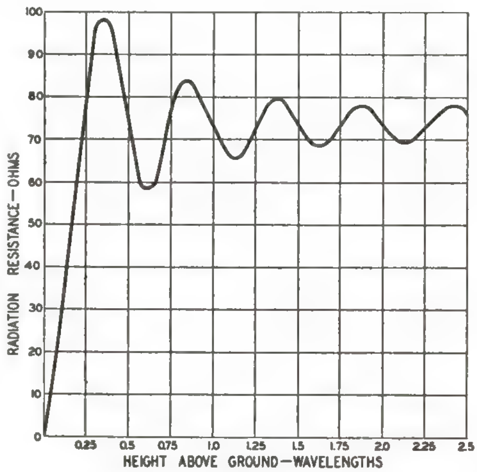


Fig. 2-45—Variation in radiation resistance of a horizontal half-wave antenna with height above perfectly conducting ground.

With other types an antennas—e.g., horizontal—at heights of a quarter wavelength or more, losses in the ground are much less important. For these types the considerable constructional problems are not justified by the possible improvement.

Ground screens will affect only the very-high-angle rays from horizontal antennas, and will not appreciably modify the effect of the earth itself at the lower radiation angles which ordinarily are used for long-distance communication.

Directive Diagrams and the Wave Angle

In the discussion of radiation patterns or directive diagrams of antennas it was brought out that such patterns always are three-dimensional affairs, but that it is difficult to show, on a plane sheet of paper, more than a cross section of the solid pattern at a time. The cross sections usually selected are those cut by a plane that contains the wire axis, and those cut by a plane perpendicular to the wire axis.

If the antenna is horizontal, the former pattern (the cross section cut by a plane containing the wire) represents the radiation pattern of the antenna when the wave leaves the antenna (or arrives at it) at zero angle of elevation above the earth. The angle of elevation—the "vertical" angle referred to in the discussion of ground effects—is usually called the wave angle. In the case of the vertical antenna the radiation pattern at zero wave angle is given by the cross section cut by the plane perpendicular to the wire axis. (In the latter case it must be remembered that the antenna itself is assumed to be merely a point at the center of the pattern, so the plane must pass through this point.)

Now with two exceptions—surface waves at low frequencies and space waves at v.h.f. and higher—the wave angle used for communication is not zero. In ionosphere transmission waves sent directly *upward* can be reflected back to earth, if the frequency is low enough; on the other hand, as pointed out earlier in this chapter, in most of the frequency range useful for ionospheric communication waves leaving at an angle of less than about 3 degrees are largely absorbed by ground losses. What we are interested in at these frequencies, then, is the directive pattern of the antenna at a wave angle that is of value in communication.

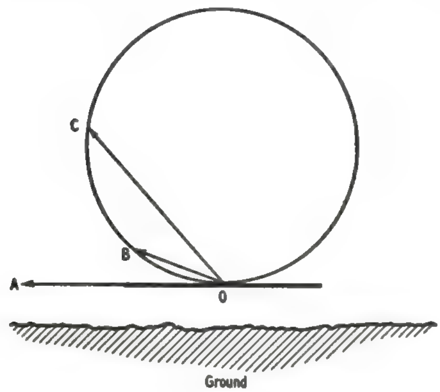


Fig. 2-46—The effective directive pattern of the antenna depends upon the angle of radiation considered. As shown by the arrows, the field strength in a given compass direction will be quite different at different vertical angles.

EFFECTIVE DIRECTIVE DIAGRAMS

The directive diagram for a wave angle of zero elevation (purely horizontal radiation) does not give an accurate indication of the directive properties of a horizontal antenna at wave angles above zero. For example, consider the half-wave dipole pattern in Fig. 2-13. It shows that there is no radiation directly in line with the antenna itself, and this is true at zero wave angle. However, if the antenna is horizontal and some wave angle other than zero is considered, it is not true at all.

The reason why will become clear on inspection of Fig. 2-46, which shows a horizontal half-wave antenna with a cross section of its free-space radiation pattern, cut by a plane that is vertical with respect to the earth and which contains the axis of the antenna conductor. (For the moment, reflections from the ground are neglected.) The lines *OA*, *OB* and *OC* all point in the same *geographical* direction (the direction in which the wire itself points), but make different angles, in the vertical plane, with the antenna. In other words, they correspond to different wave angles or angles of radiation, with all three rays aimed along the same line on the earth's surface. So far as compass directions are concerned, *all three waves are leaving the end of the antenna.*

The purely horizontal wave *OA* has zero amplitude, but at a somewhat higher angle corresponding to the line *OB* the field strength is appreciable. At a still higher angle corresponding to the line *OC* the field strength is still greater. In this particular pattern, the higher the wave angle the greater the field strength in the same compass direction. It should be obvious that it is necessary, in plotting a directive diagram that purports to show the behavior of the antenna in different compass directions, to specify the angle of radiation for which the diagram applies. When the antenna is horizontal the shape of the diagram will be altered considerably as the wave angle is changed.

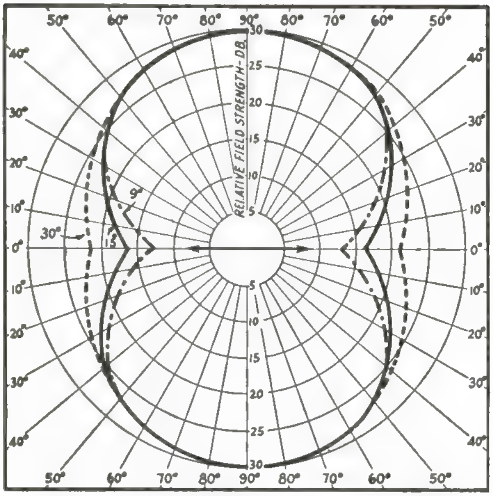


Fig. 2-47—Directive patterns of a horizontal half-wave antenna at three radiation angles, 9, 15 (solid line) and 30 degrees. The direction of the antenna itself is shown by the arrow. These patterns are plotted to a 30-db. scale, which is about proportional to signal strength as determined by ear. If 30 db. represents an S9 signal, 0 on the scale will be about S1. All three patterns are plotted to the same maximum, but the actual amplitudes at the various angles will depend upon the antenna height, as described in the text. The patterns shown here show only the shape of the directive diagram as the angle is varied.

As described in Chapter One, the wave angles that are useful depend on two things—the distance over which communication is to be carried on, and the height of the ionosphere layer that does the reflecting. Whether the E or F_2 layer (or a combination of the two) will be used depends on the operating frequency, the time of day, season, and the sunspot cycle. The same half-wave antenna, operating on the same frequency, may be almost nondirectional for distances of a few hundred miles but will give substantially better results broadside than off its ends at distances of the order of 1000-1200 miles during the day when transmission is by the E layer. In the evening, when the F layer takes over, the directivity may be fairly well marked at long distances and not at all pronounced at 1000 miles or less. From this it might seem that it would be impossible to predict the directivity of an antenna without all sorts of qualifications. However, it is possible to get a very good idea of the directivity by choosing a few angles that, on the average, are representative for different types of work. With patterns for such angles available it is fairly simple to interpolate for intermediate angles. Combined with some knowledge of the behavior of the ionosphere, a fairly good estimate of the directive characteristics of a particular antenna can be made for the particular time of day and distance of interest.

In the directive patterns given in Figs. 2-47 to 2-53, inclusive, the wave angles considered are 9, 15 and 30 degrees. These represent, respectively, the median values of a range of angles that have been found to be effective for communication at 28, 21, 14 and 7 Mc. Because of the variable nature of ionosphere propagation the patterns should not be considered to be more than general guides to the sort of directivity to be expected.

In the directive patterns of Figs. 2-47 to 2-53 the relative field strength has been plotted in decibels. This makes the patterns more representative of the effect produced than is the case when the relative intensity is plotted in either voltage or power. Since one "S" point on the signal-strength scale is roughly 5 or 6 db., it is easy to get an approximte idea of the operation of the antenna. For example, off the ends of a half-wave antenna the signal can be expected to be "down" between 2 and 3 "S" points compared with its strength at right angles or broadside to the antenna, at a wave angle of 15 degrees. This would be fairly representative of its performance on 14 Mc. at distances of 500 miles or more. With a wave angle of 30 degrees the signal off the ends would be down only 1 to 2 "S" points, while at an angle of 9 degrees it would be down 3 to 4 "S" points. Since high wave angles become less useful as the frequency is increased, this illustrates the importance of running the antenna wire in the proper direction if best results are wanted in a particular direction at the higher frequencies.

Height Above Ground

The *shapes* of the directive patterns given in Figs. 2-47 to 2-53 are not affected by the height of the antenna above the ground. However, the *amplitude* relationships between the patterns of

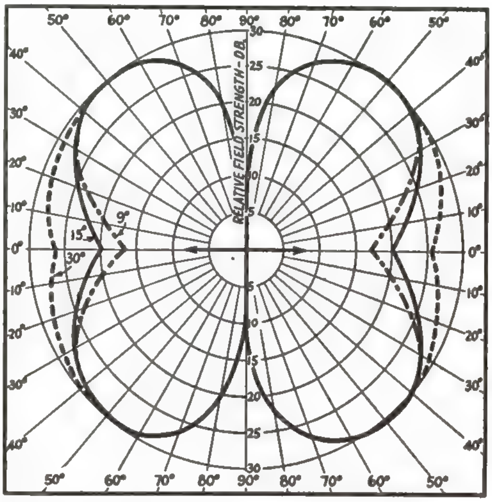


Fig. 2-48—The horizontal patterns for a one-wavelength antenna at vertical angles of 9, 15 and 30 degrees.

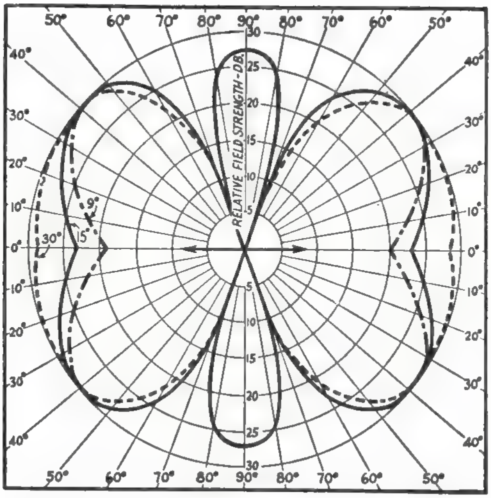


Fig. 2-49—The horizontal patterns for a 1½-wavelength antenna at vertical angles of 9, 15 and 30 degrees.

a given antenna for various wave angles are modified by the height. In the figures as given, the scale is such that the same field intensity is assumed in the direction of maximum radiation, regardless of the wave angle. To make best use of the patterns the effect of the ground-reflection factor should be included.

Take the horizontal half-wave antenna shown in Fig. 2-47 as an example, and assume that the antenna is a half wavelength above perfectly-conducting ground. The graph of the ground-reflection factor for this height is given in Fig. 2-29. For angles of 9, 15 and 30 degrees the values of the factor as read from the curve are 1.0, 1.5 and 2.0, respectively. These factors are applied to field strength. For convenience, take the

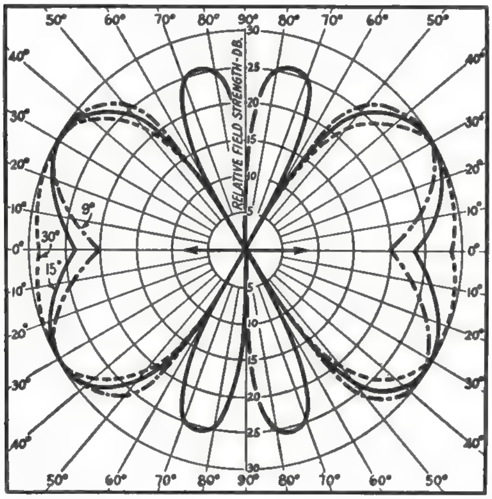


Fig. 2-50—The horizontal patterns for a 2-wavelength antenna at vertical angles of 9, 15 and 30 degrees.

9-degree angle as a reference. Then at a wave angle of 15 degrees the field strength will be 1.5 times the field strength at 9 degrees, *in any compass direction*, and at a wave angle of 30 degrees will be 2.0 times the field strength at 9 degrees, in any compass direction. Since the factors apply to field strength (voltage) the ratios just obtained may be converted to decibels by using the voltage curve in Fig. 2-22. A voltage ratio of 1.5 corresponds to 3.5 db., and a ratio of 2.0 corresponds to 6 db. Hence at an angle of 15 degrees the radiation in any direction is 3.5 db. above the radiation at 9 degrees, and at a wave angle of 30 degrees it is 6 db. above. To put it another way, at a wave angle of 30 degrees the antenna is about an "S" point better than it is at 9 degrees. There is about a half "S"-point difference between 9 and 15 degrees, and between 15 and 30 degrees. If we wanted, we

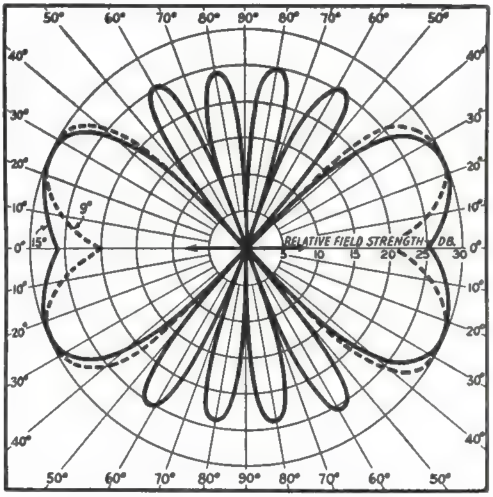


Fig. 2-51—The horizontal patterns for a 3-wavelength antenna at vertical angles of 9 and 15 degrees.

could add 3.5 db. to every point on the 15-degree graph in Fig. 2-47, and 6 db. to every point on the 30-degree graph, and thus show graphically the comparison in amplitude as well as shape of the directive pattern at the three angles. This has been done in the graph of Fig. 2-54. However, it is generally unnecessary to take the trouble to draw separate graphs because it is so easy to add or subtract the requisite number of decibels as based on the appropriate ground-reflection factor.

It should be emphasized again that the patterns are based on idealized conditions not realized over actual ground. Nevertheless, they are useful in indicating about what *order* of effect to expect.

Using the Patterns

Directive patterns can be of considerable help in solving practical problems in the choice and location of antennas, particularly in cases where

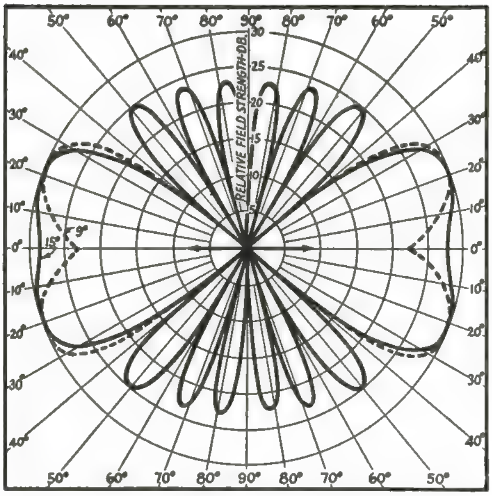


Fig. 2-52—The horizontal patterns for a 4-wavelength antenna at vertical angles of 9 and 15 degrees.

a simple type of antenna (such as the antennas discussed in this chapter) has to be used. While it is not to be expected that antenna performance will conform to the theoretical with the same exactness that you would expect Ohm's Law calculations to work out, the results averaged over a period of time will be sufficiently close to the predictions to make a little preliminary estimating worth while.

Here is one example: Suppose that a clear space of about 70 feet is available between two supports that will hold the antenna about 35 feet above ground. The operating frequency is to be 28 Mc. and the positions of the supports is such that the antenna will run west of north by 10 degrees. The principal direction of transmission is to be 35 degrees east of north, but there is another area in the general direction 15 degrees south of west that it is also hoped to cover as well as possible. The situation is shown in Fig. 2-55 (in this figure the last direction is shown with reference to the north-south line).

In the available space, it is possible to erect antennas ½, 1, 1½ or 2 wavelengths long. Since the supports are at fixed height, the ground-reflection factor will be the same for all the possible antennas and so may be left out of the estimates. The principal direction is 45 degrees off the line of the antenna and the secondary direction is 85 degrees off. For simple antennas such as these the directive patterns are symmetrical about the wire axis and so we do not have to worry about whether the angles lie east or west of the antenna.

Since the frequency is 28 Mc. the 9-degree patterns should be used. From Fig. 2-47 we see that the relative amplitudes at 45 and 85 degrees are 27 and 30 db., respectively, for the half-wave antenna. From Fig. 2-48 the corresponding amplitudes are 30 and 17 db.; from Fig. 2-49 the amplitudes are 28 and 27 db.; and from Fig. 2-50 the amplitudes are 28 and 20 db. To these amplitudes we should add the gains realized by harmonic operation as given in Fig. 2-23. These are, for the ½-, 1-, 1½- and 2-wavelength antennas, respectively, 1, 1.1, 1.2 and 1.3 in power. Converted to decibels by using Fig. 2-22 they are 0, 0.5, 0.8 and 1 db. respectively. They are small enough to be less important than the probable error in reading the charts, but will be included for the sake of completeness. Arranging the information in table form gives

	Antenna Length in Wavelengths			
	0.5	*1*	*1.5*	*2*
Relative intensity at 45 degrees, db.	27	30	28	28
Gain from harmonic operation, db.	0	0.5	0.8	1
Total	*27*	*30.5*	*28.8*	*29*
Relative intensity at 85 degrees, db.	30	17	27	20
Gain from harmonic operation, db.	0	0.5	0.8	1
Total	*30*	*17.5*	*27.8*	*21*

It is seen that either a 1-wavelength or 1½-wavelength antenna will give the best results in the principal direction, but that neither is as good as a half-wave antenna in the secondary direction. In a case such as this, the best all-around results would be obtained by using *two* antennas, since there is room to string them end to end. A good combination, for example, would be a 1-wavelength and ½-wavelength antenna, arranged with a little space between the ends so the coupling between the pair of antennas is substantially reduced.

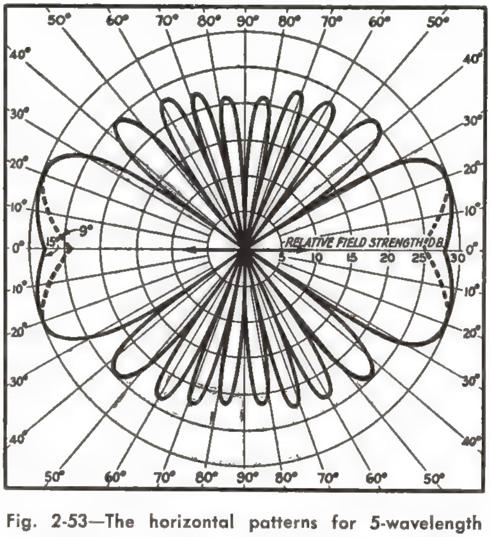


Fig. 2-53—The horizontal patterns for 5-wavelength antenna at vertical angles of 9 and 15 degrees.

Another example: Space is available to erect a 4-wavelength antenna at a height of 30 feet, for operation on 28 Mc., and it is possible to orient the antenna so that its major lobe will point in the direction of transmission desired. Alternatively, a self-supporting half-wave dipole could be erected at a height of 45 feet and oriented so that its maximum radiation would be in the desired direction. Which antenna is likely to be the better one?

From Fig. 2-23, the power gain of a 4-wavelength antenna over a dipole is 2.1, and from Fig. 2-22 this power ratio corresponds to a gain of 3 db. At 28 Mc. the length of a half wavelength is $492/28 = 17.6$ feet, so one wavelength is 35 feet, near enough. At 30 feet the height in wavelengths is $30/35 = 0.86\lambda$, and at 45 feet it is $45/35 = 1.3\lambda$. (It is not necessary to carry out the calculations to more than two significant figures because the height above the effective ground plane is not known. As explained earlier, the "effective" height would tend to be higher than the actual height.) A height of 0.85λ is near enough to $\frac{7}{8}\lambda$ to permit us to use Fig. 2-32, in which the ground-reflection factor is shown as 1.5 at a wave angle of 9 degrees. From Fig. 2-22 the corresponding figure is 3.5 db. A height of 1.3λ is slightly over $1\frac{1}{4}\lambda$, and inspecting Figs. 2-34 and 2-35 shows that the reflection factor therefore will lie in the vicinity of 1.9, or about 5.5 db. Putting these figures into table form, we have

	Dipole	*4λ Antenna*
Relative intensity of main lobe, db.	30	30
Gain from harmonic operation, db.	0	3
Ground-reflection factor, db.	5.5	3.5
Total	*35.5*	*36.5*

The difference, 1 db., is in favor of the 4λ antenna, but for all practical purposes the two antennas could be considered equally good in the desired direction; the additional height of the dipole is just about sufficient to overcome the gain of the harmonic antenna. The choice therefore could be based purely on other considerations such as convenience in erection, the fact that the dipole antenna will be effective over a wider horizontal angle than the harmonic antenna, and so on.

Once again it must be emphasized that calculations such as these should not be taken *too* literally. Too many factors, particularly the behavior of the ground, are unknown. The calculations such as these should not be taken *too* determining the type of antenna that, in all probability, will best meet the required working conditions.

RADIATION RESISTANCE AND GAIN

The field strength produced at a distant point by a given antenna system is directly proportional to the current flowing in the antenna. In turn, the amount of current that will flow, when a fixed amount of power is applied, will be inversely proportional to the square root of the radiation resistance. Lowering the radiation resistance will increase the field strength and raising the radiation resistance will decrease it. This is not to be interpreted broadly as meaning that a low value of radiation resistance is good and a high value is bad, regardless of circumstances, because that is far from the actual case. What it means is that, with an antenna of given dimensions, a change that reduces the radiation resistance *in the right way* will be accompanied by a change in the directive pattern that in turn

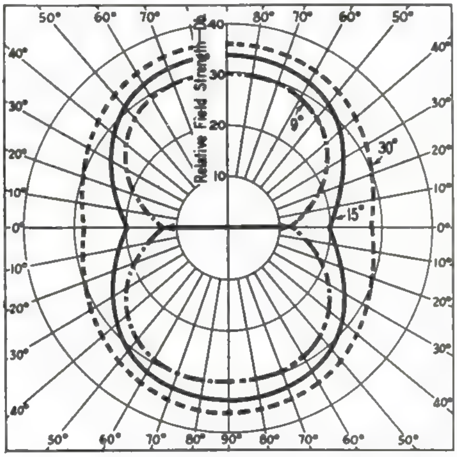


Fig. 2-54—These diagrams compare the amplitudes of radiation at wave angles of 9, 15 and 30 degrees from a horizontal half-wave antenna when the height is ½ wavelength.

will increase the field strength in some directions at the expense of reduced field strength in other directions. This principle is used in certain types of directive systems described in detail in a later chapter.

The shape of the directional pattern in the vertical plane is, as previously described, modified by the height of the antenna above ground. The effect of height on radiation resistance has been shown in Fig. 2-45 for the horizontal half-wave dipole. The plots of ground-reflection factors shown in Figs. 2-26 to 2-37, inclusive, show the actual shape of the pattern of such an antenna in the vertical plane at right angles to the wire. That is, they show the variation in intensity with wave angle in the direction broadside to the antenna. In an approximate way, the radiation resistance is smaller as the area of the pattern is less, as can be seen by comparing the ground-reflection patterns with the curve of Fig. 2-45.

Varying the height of a horizontal half-wave antenna while the power input is held constant will cause the current in the antenna to vary as its radiation resistance changes. Under the ideal-

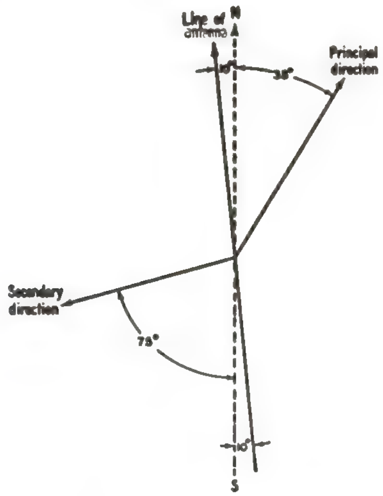


Fig. 2-55—Example discussed in the text.

ized conditions represented in Fig. 2-45 (an infinitely-thin conductor over perfectly-conducting ground) the field intensity at the optimum wave angle for each height will vary as shown in Fig. 2-56. In this figure the relative field intensity is expresed in decibels, using the field when the radiation resistance is 73 ohms as a reference (0 db.). From this cause alone, there is a gain of about 1 db. when the antenna height is ⅝ wavelength as compared with either ½ or ¾ wavelength. The gain or loss from the change in radiation resistance should be combined with the reflection factor for the particular wave angle and antenna height considered, in judging the over-all effect of height on performance.

For example, Fig. 2-57 shows the reflection factor, plotted in decibels, for a wave angle of 15 degrees (solid curve). This curve is based on data from Figs. 2-26 to 2-37, inclusive. Taken alone it would indicate that a height of slightly less than 1 wavelength is optimum for this wave angle. However, when the values taken from the curve of Fig. 2-56 are added the broken curve results. Because of the change in radiation resistance, there is a maximum near a height of ⅝ wavelength that is very nearly as good as the

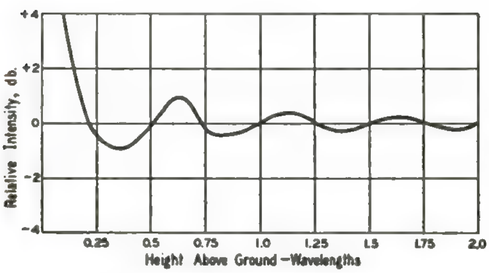


Fig. 2-56—Gain or loss in decibels because of change in antenna current with radiation resistance, for fixed power input. Perfectly-conducting ground is assumed.

next maximum at a height of 1 to 1¼ wavelength. The change in radiation resistance also has the effect of steepening the curve at the lower heights and flattening it in the optimum region. Thus it would be expected that, for this wave angle, increasing the height of a half-wave dipole is very much worth while up to about ⅝ wavelength, but that further increases would not result in any material improvement. At 14 Mc., where a 15-degree wave angle is taken to be average, ⅝ wavelength is about 45 feet.

There is, of course, some difficulty in applying the information obtained in this fashion because of the uncertainty as to just where the ground plane is. One possibility, if the antenna can be raised and lowered conveniently, is to measure the current in it while changing its height, keeping the power input constant. Starting with low heights, the current should first go through a minimum (at a theoretical height of about ⅜ wavelength) and then increase to a maximum as the height is increased. The height at which this maximum is obtained is the optimum.

It should be kept in mind that no one wave angle does all the work. Designing for optimum results under average conditions does not mean that best results will be secured for all types of

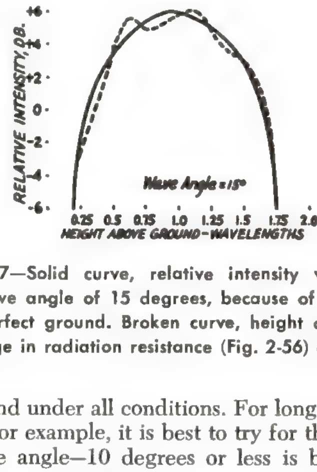


Fig. 2-57—Solid curve, relative intensity vs. height at a wave angle of 15 degrees, because of reflection from perfect ground. Broken curve, height and effect of change in radiation resistance (Fig. 2-56) combined.

work and under all conditions. For long-distance work, for example, it is best to try for the lowest possible angle—10 degrees or less is better for multihop propagation at 14 Mc., for example. However, an antenna that radiates well at such low angles may not be as good for work over shorter distances as one having a broader lobe in the vertical plane.

The effect of radiation resistance is somewhat more marked at the lower frequencies. To cover a distance of 200 miles at night (*F*-layer propagation) requires a wave angle of 60 degrees. As shown by the patterns of Figs. 2-26 to 2-28, the optimum antenna height for this wave angle is ¼ wavelength. However, it is in the region below ¼ wavelength that the radiation resistance decreases most rapidly. At a height of ⅛ wavelength there is a gain of 3.5 db. over a height of ¼ wavelength because of lowered radiation resistance. To offset this, the ground-reflection factor for a wave angle of 60 degrees is about 1.25

at ⅛ wavelength (Fig. 2-26) as compared with 2.0 for ¼ wavelength; this is a loss of 4 db. There is thus a difference of only ½ db., which is not observable, between ⅛ and ¼ wavelength. At 3.5 Mc. this is a considerable difference in actual height, since ⅛λ is about 35 feet and ¼λ is about 70 feet. For *short*-distance work the cost of the supports required for the greater height would not be justified.

Information on the variation in radiation resistance with height for other types of antenna than the half-wave dipole is not readily available. A harmonic antenna can be expected to show such variations, but in general an antenna system that tends to minimize the radiation directly toward the ground under the antenna can be expected to have a lesser order of variation in radiation resistance with height than is the case with the half-wave dipole.

VERTICAL DIRECTIVITY PATTERNS

It was explained in the preceding section that the directive patterns of Figs. 2-47 to 2-53, inclusive, show the relative intensity of radiation in different compass directions for each wave angle selected, but do not attempt to show the amplitude relationship between the wave angles. This is because the intensity at different wave angles varies with the height of the antenna above ground, and an extremely large number of diagrams would be needed to represent the range of heights and lengths of antennas encountered in practice. The information on relative intensity at different wave angles is easily secured from the ground-reflection charts.

However, it is helpful in forming a picture of the operation of antennas to use a form of representation in which the *vertical* directional characteristic is shown for different heights. Inasmuch as we are still confronted by a three-dimensional pattern, it is only possible to do this for selected vertical planes oriented in various directions with respect to the antenna axis. In the case of the horizontal half-wave dipole a

Vertical-Plane Radiation Patterns of Horizontal Half-Wave Antennas Above Perfectly-Conducting Ground

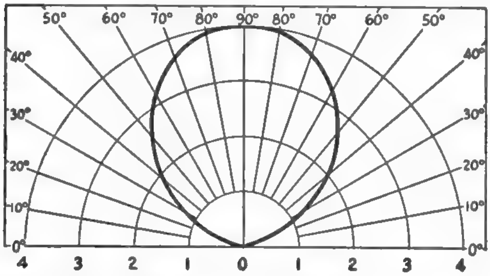


Fig. 2-58 — In direction of wire; height ¼ wavelength.

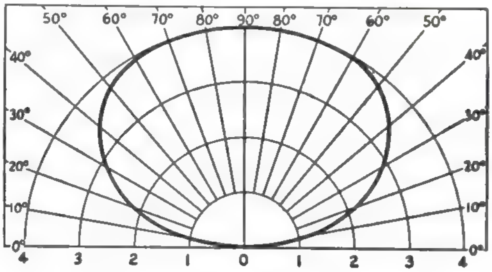


Fig. 2-59 — At right angles to wire; height ¼ wavelength.

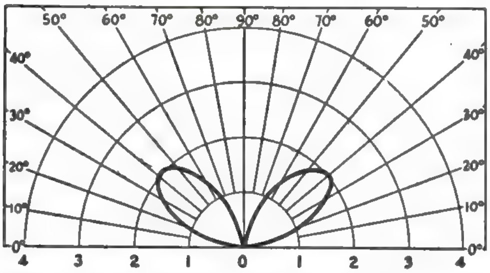


Fig. 2-60 — In direction of wire; height ½ wavelength.

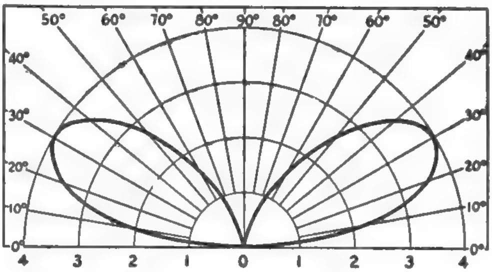


Fig. 2-61 — At right angles to wire; height ½ wavelength.

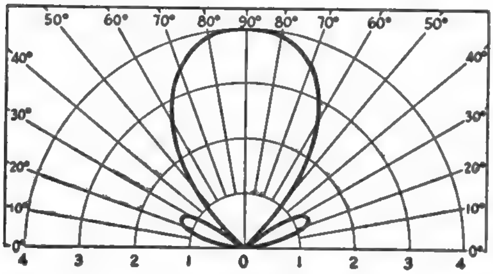


Fig. 2-62 — In direction of wire; height ¾ wavelength.

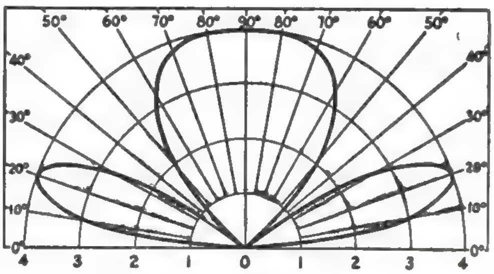


Fig. 2-63 — At right angles to wire; height ¾ wavelength.

plane running in a direction along the axis and another broadside to the antenna will give a good deal of information.

A series of such patterns for a horizontal half-wave dipole at different heights is given in Figs. 2-58 to 2-69, inclusive. The scale is simply an arbitrary one in which the length of a radius drawn from the origin to any point on the graph is proportional to field strength (voltage). The reduction in field strength off the ends of the wire at the lower angles, as compared with the broadside field strength, is quite apparent. It is also apparent that, at some heights, the high-angle radiation off the ends is nearly as great as the broadside radiation.

In vertical planes making some angle intermediate between 0 and 90 degrees with the wire axis the pattern will have a shape intermediate between the two planes shown. By visualizing a smooth transition from the end-on pattern to the broadside pattern as the horizontal angle is varied from 0 to 90 degrees a fairly-good mental picture of the actual solid pattern can be formed.

In the case of a vertical half-wave dipole, the horizontal directional pattern is simply a circle at any wave angle (although the actual field strength will vary, at the different wave angles, with the height above ground). Hence one vertical pattern is sufficient to give complete information, for a selected antenna height, about the antenna in any direction with respect to the wire. A series of such patterns is given in Figs. 2-70 to 2-73, inclusive. These patterns are formed by multiplying one lobe of the free-space pattern of a half-wave dipole by the ground-reflection factor that applies at each wave angle for the antenna height selected, to obtain the resultant relative field strength at each wave angle. The solid pattern in each case is formed by rotating the plane pattern about the 90-degree axis of the graph.

The effect of ground losses at high frequencies is simulated by the broken curves at the very low wave angles. In other respects the

Vertical-Plane Radiation Patterns of Horizontal Half-Wave Antennas Above Perfectly-Conducting Ground

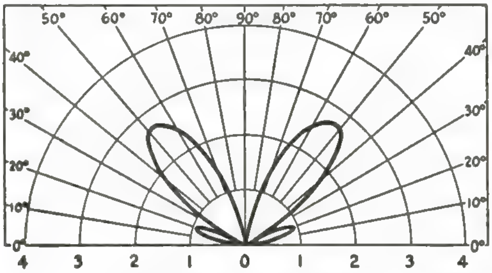


Fig. 2-64 — In direction of wire; height 1 wavelength.

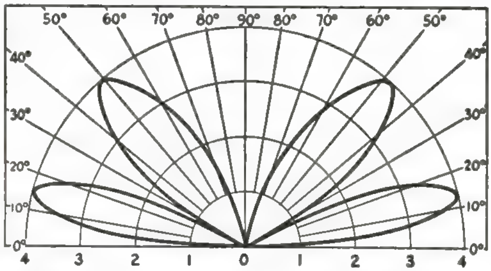


Fig. 2-65 — At right angles to wire; height 1 wavelength.

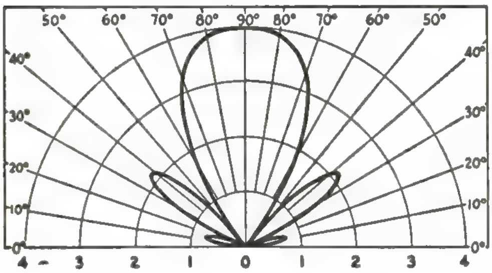


Fig. 2-66 — In direction of wire; height 1¼ wavelength.

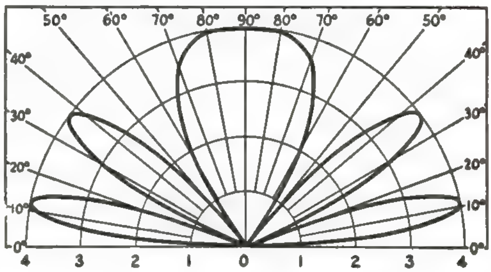


Fig. 2-67 — At right angles to wire; height 1¼ wavelength.

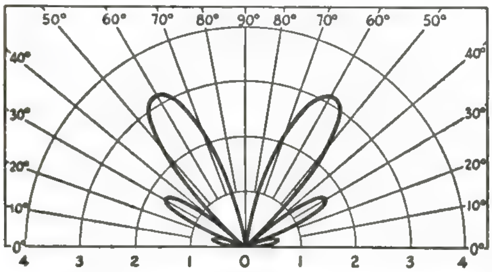


Fig. 2-68 — In direction of wire; height 1½ wavelength.

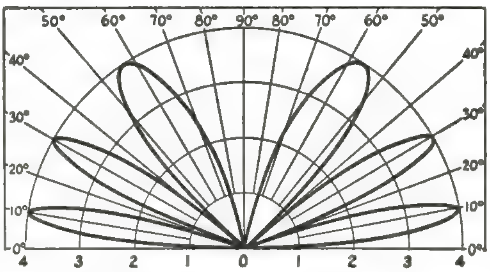


Fig. 2-69 — At right angles to wire; height 1½ wavelength.

curves are based on the assumption that the antenna is erected over perfectly-conducting ground.

SOME PRACTICAL CONSIDERATIONS

At the risk of being repetitious, we must state again that the results from a practical antenna cannot be expected to be exactly according to the theoretical performance outlined in this chapter. The theory that leads to the impedances, radiation patterns, and power-gain figures discussed is necessarily based on idealized assumptions that cannot be exactly realized, although they may be approached, in practice.

The effect of imperfectly-conducting earth has been mentioned several times. It will cause the actual radiation resistance of an antenna to differ somewhat from the theoretical figure at a given height. In addition, there is the effect of the length/diameter ratio of the conductor to be considered. Nevertheless, the theoretical figure will approximate the actual radiation resistance closely enough for most practical work. The value of radiation resistance is of principal importance in determining the proper method for feeding power to the antenna through a transmission line, and a variation of 10 or even 20 per cent will not be serious. Adjustments can easily be made to compensate for the discrepancy between practice and theory.

So far as radiation patterns are concerned, the effect of imperfect earth is to decrease the amplitude of the reflected ray and to introduce some phase shift on reflection. The phase shift is generally small with horizontal polarization. Both effects combine to make the maximum reflection factor somewhat less than 2, and to prevent complete cancellation of radiation in the nulls in the theoretical patterns. There may also be a slight change in the wave angle at which maximum reinforcement occurs, as a result of the phase shift. The effect of ground losses on very low angles already has been emphasized.

Aside from ground effects, the theoretical patterns of the antennas discussed are developed on the basis of sinusoidal distribution of current along the antenna, and on the assumption (in harmonic antennas) that the value of the current is the same at every current loop. Neither is strictly true. In particular, the current in a long harmonic antenna is not the same at every loop because some energy is lost all along the antenna by radiation. This affects both the current going out and the current returning after reflection at the far end of the antenna. The result is that the radiation pattern does not have the perfect symmetry indicated in the drawings in this chapter. The lobes pointing away from the end at which the antenna is fed are tilted somewhat toward the direction of the antenna wire, and the lobes pointing toward the fed end are tilted away from the wire. The latter also have less amplitude than the former. Typical measured patterns are shown in Fig. 2-74. There is even a tilt to the pattern of a half-wave antenna when it is fed at one end; however, when such an antenna is fed at the center the pattern is symmetrical.

Finally, the effect of near-by conductors and dielectrics cannot, of necessity, be included in

Vertical-Plane Radiation Patterns of Horizontal Half-Wave Antennas Above Perfectly-Conducting Ground The height is that of the center of the antenna. Dotted lines indicate approximate effect of attenuation of the very low-angle radiation because of ground losses.

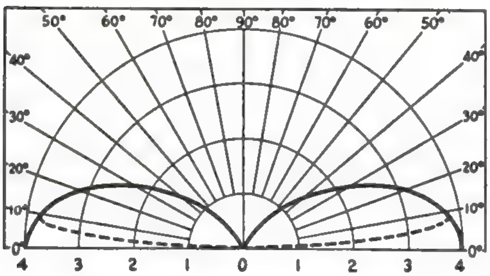


Fig. 2-70 — Height ¼ wavelength.

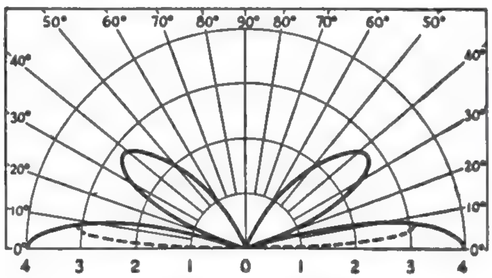


Fig. 2-72 — Height ¾ wavelength.

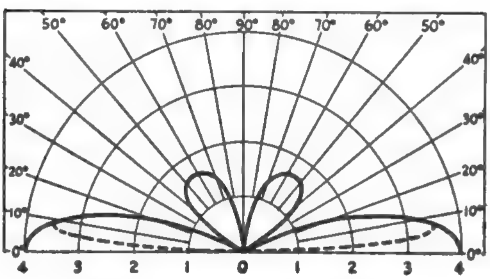


Fig. 2-71 — Height ½ wavelength.

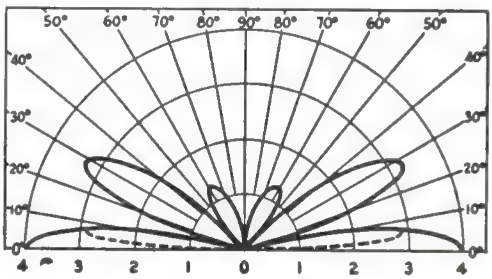


Fig. 2-73 — Height 1 wavelength.

the theoretical patterns. Conductors such as power and telephone lines, house wiring, piping, etc., close to the antenna can cause considerable distortion of the pattern if currents of appreciable magnitude are induced in them. Under similar conditions they can also have a marked effect on the radiation resistance. Poor dielectrics such as green foliage near the antenna can introduce loss, and may make a noticeable difference between summer and winter performance.

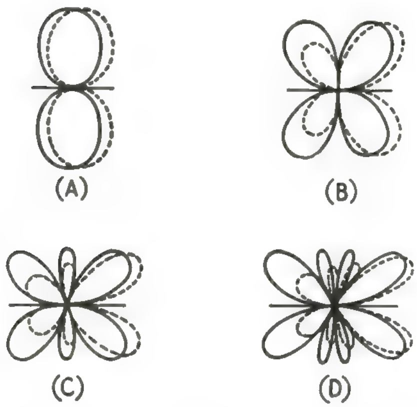


Fig. 2-74 — The effect of feeding an antenna at the end is to cause a tilt to the directional pattern, as shown by these experimentally-determined patterns. A, half wave; B, 1 wavelength; C, 1½ wavelengths; D, 2 wavelengths. Solid patterns are theoretical, dotted patterns experimental. In each case the antenna is fed from the left-hand end.

The directional effects of an antenna will conform more closely to theory if the antenna is located in a clear space, at least a half wavelength from anything that might affect its properties. In cities, it may be difficult to find such a space at low frequencies. The worst condition arises when near-by wires or piping happen to be resonant, or nearly so, at the operating frequency. Such resonances often can be destroyed by bonding pipes or BX coverings at trial points, checking with a crystal-detector wavemeter to determine the measures necessary to reduce the induced current. Metal masts or guy wires can cause distortion of the pattern unless detuned by grounding or by breaking up the wires with insulators. However, masts and guy wires usually have relatively little effect on the performance of horizontal antennas because, being vertical or nearly so, they do not pick up much energy from a horizontally-polarized wave. In considering near-by conductors, too, the transmission line that feeds the antenna should not be overlooked. Under some conditions that are rather typical with amateur antennas currents will be induced in the line by the antenna, leading to some undesirable effects. This is considered in Chapter Three.

Special Antenna Types

The underlying principles of antenna operation have been discussed in this chapter in terms of the half-wave dipole, which is the elementary form from which more elaborate antenna systems are built. However, there are other types of antennas that find some application in amateur work, particularly when space limitations do not permit using a full-sized dipole. These include, principally, grounded antennas and loops.

THE GROUNDED ANTENNA

In cases where vertical polarization is required —for example, when a low wave angle is desired at frequencies below 4 Mc.—the antenna must be vertical. At low frequencies the height of a vertical half-wave antenna would be beyond the constructional reach of most amateurs. A 3.5-Mc. vertical half wave would be 133 feet high, for instance.

However, if the lower end of the antenna is grounded it need be only a quarter wave high to resonate at the same frequency as an ungrounded half-wave antenna. The operation can be understood when it is remembered that ground having high conductivity acts as an electrical mirror, and the missing half of the antenna is supplied by the mirror image. This is shown in Fig. 2-75.

The directional characteristic of a grounded quarter-wave antenna will be the same as that of a half-wave antenna in free space. Thus a vertical grounded quarter-wave antenna will have a circular radiation pattern in the horizontal plane. In the vertical plane the radiation will decrease from maximum along the ground to zero directly overhead.

The grounded antenna may be much smaller than a quarter-wave and still be made resonant by "loading" it with inductance at the base, as in Fig. 2-76 at B and C. By adjusting the inductance of the loading coil even very short wires can be tuned to resonance.

The current along a grounded quarter-wave vertical wire varies practically sinusoidally, as is the case with a half-wave wire, and is highest at

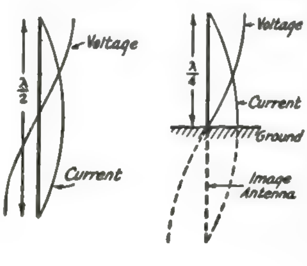


Fig. 2-75 — The half-wave antenna and its grounded quarter-wave counterpart. The missing quarter-wavelength can be considered to be supplied by the image in ground of good conductivity.

the ground connection. The r.f. voltage, however, is highest at the open end and minimum at the ground. The current and voltage distribution are shown in Fig. 2-76A. When the antenna is shorter than a quarter wave but is loaded to resonance, the current and voltage distribution are part sine waves along the antenna wire. If the loading coil is substantially free from distributed capacity, the voltage across it will increase uniformly from minimum at the ground, as shown at B and C, while the current will be the same throughout.

Extremely short antennas are used, of necessity, in mobile work on the lower frequencies such as the 3.5-Mc. band. These may be "base-loaded" as shown at B and C in Fig. 2-76, but there is a small advantage to be realized by placing the loading coil at the center of the antenna. In neither case, however, is the current uniform throughout the coil, since the inductance required is so large that the coil tends to act like a linear circuit rather than like a "lumped" inductance.

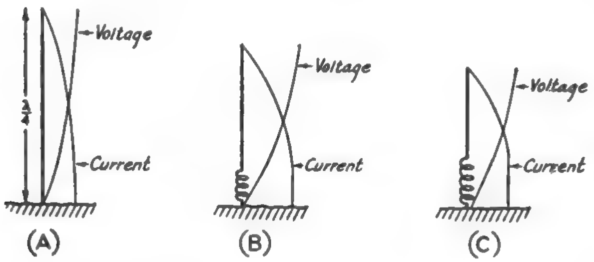


Fig. 2-76 — Current and voltage distribution on a grounded quarter-wave antenna (A) and on successively shorter antennas loaded to resonate at the same frequency.

If the antenna height is greater than a quarter wavelength the antenna shows inductive reactance at its terminals and can be tuned to resonance by means of a capacitance of the proper value. This is shown in Fig. 2-77. As the length is increased progressively from ¼ to ½ wavelength the current loop moves up the antenna, always being at a point ¼ wavelength from the top. When the height is ½ wavelength the current distribution is as shown at B in Fig. 2-77. There is a voltage loop (current node) at the base, and power can be applied to the antenna through a parallel-tuned circuit, resonant to the same frequency as the antenna, as shown in the figure.

Up to a little more than ½ wavelength, increasing the height compresses the directive pattern in the vertical plane; this results in an increase in field strength, for a given power input, at the very low radiation angles. The theoretical improvement is about 2 db. for a half-wave antenna as compared with a quarter-wave antenna.

Radiation Resistance

The radiation resistance of a grounded vertical antenna, as measured between the base of the antenna and ground, varies as shown in Fig. 2-78 as a function of the antenna height. The word "height" as used in this connection has the same meaning as "length" as applied to a horizontal antenna. This curve is for an antenna based on (but not directly connected to) ground of perfect conductivity. The height is given in electrical degrees, the 60-135 degree range shown corresponding to heights from ⅙ to ⅜ wavelength. The antenna is approximately self-resonant at a height of 90 degrees (¼ wavelength). The actual resonant length will be somewhat less because of the length/diameter ratio mentioned earlier.

In the range of heights covered by Fig. 2-78 the radiation resistance is practically independent of the length/diameter ratio. At greater heights the length/diameter ratio is important in determining the actual value of radiation resistance. At a height of ½ wavelength the radiation resistance may be as high as several thousand ohms.

The variation in radiation resistance with heights below 60 degrees is shown in Fig. 2-79. The values in this range are very low.

A very approximate curve of reactance *vs.* height is given in Fig. 2-80. The actual reactance will depend on the length/diameter ratio, so this curve should be used only as a rough guide. It is based on a ratio of about 1000 to 1. Thicker antennas can be expected to show lower reactance at a given height, and thinner antennas should show more. At heights below and above the range covered by the curve, larger reactance values will be encountered, except for heights in the vicinity of ½ wavelength. In this region the reactance decreases, becoming zero when the antenna is resonant.

Efficiency

The **efficiency** of the antenna is the ratio of the radiation resistance to the total resistance of the system. The total resistance includes radiation resistance, resistance in conductors and dielectrics, including the resistance of loading coils if used, and the resistance of the grounding sys-

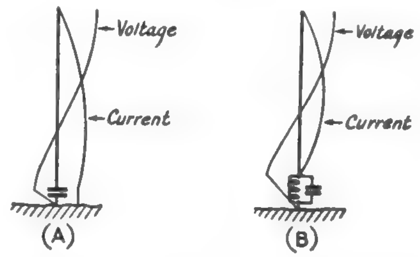


Fig. 2-77 — Current and voltage distribution on grounded antennas longer than ¼ wavelength. (A), between ¼ and ⅜ wave, approximately; (B) half-wave.

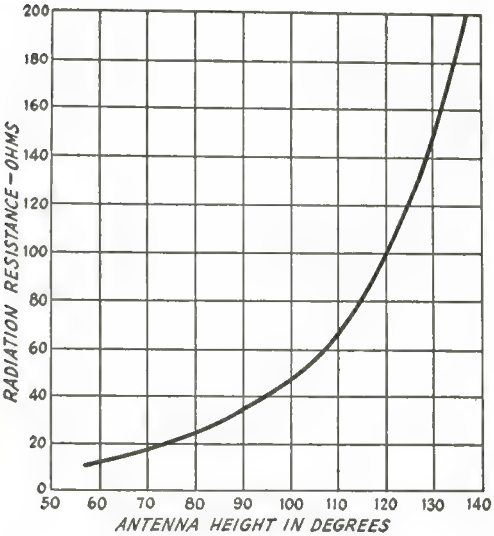


Fig. 2-78 — Radiation resistance vs. antenna height in degrees, for a vertical antenna over perfectly-conducting ground, or over a highly-conducting ground plane. This curve also may be used for center-fed antennas (in free space) by multiplying the radiation resistance by two; the height in this case is half the actual antenna length.

tem, usually referred to as "ground resistance."

It was stated earlier in this chapter that a half-wave antenna operates at very high efficiency because the conductor resistance is negligible compared with the radiation resistance. In the case of the grounded antenna the ground resistance usually is not negligible, and if the antenna is short (compared with a quarter wavelength) the resistance of the necessary loading coil may become appreciable. To attain an efficiency comparable with that of a half-wave antenna, in a grounded antenna having a height of ¼ wavelength or less, great care must be used to reduce both ground resistance and the resistance of any required loading inductors. Without a fairly elaborate grounding system, the efficiency is not likely to exceed 50 per cent and may be much less, particularly at heights below ¼ wavelength.

Grounding Systems

The ideal grounding system for a vertical grounded antenna would consist of about 120 wires, each at least ½ wavelength long, extending radially from the base of the antenna and spaced equally around a circle. Such a system is the practical equivalent of perfectly-conducting ground and has negligible resistance. The wires can either be laid directly on the surface of the ground or buried a few inches below.

Such a system would be practical in an amateur installation in very few cases. Unfortunately, the resistance increases rapidly when the number of radials is reduced, and if at all possible at least 15 radials should be used. Experimental measurements have shown that even with this number the resistance is such as to decrease the antenna efficiency to about 50 per cent at a height of ¼ wavelength.

It has also been found that as the number of radials is reduced the length required for optimum results with a particular number of radials also decreases; in other words, if only a small number of radials can be used there is no point in extending them out a half wavelength. With 15 radials, for example, a length of ⅛ wavelength is sufficient. With as few as two radials the length is almost unimportant, but the efficiency of a quarter-wave antenna with such a grounding system is only about 25 per cent. It is considerably lower with shorter antennas.

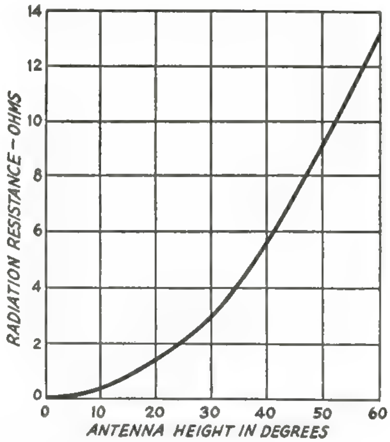


Fig. 2-79 — Same as Fig. 2-78, for heights below 60 degrees.

In general, a large number of radials, even though some or all of them have to be short, is preferable to a few long radials. The conductor size is relatively unimportant, No. 12 or No. 14 copper wire being suitable.

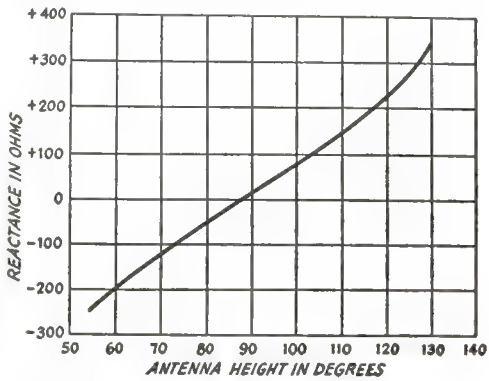


Fig. 2-80. — Approximate reactance of a vertical antenna over perfectly-conducting ground and having a length/diameter ratio of about 1000. Actual values will vary considerably with length/diameter ratio. The remarks under Fig. 2-78 also apply to this curve.

The measurement of ground resistance at the operating frequency is difficult. The power loss in the ground depends on the current concentration near the base of the antenna, and this depends on the antenna height. Typical values for small radial systems (15 or less) have been measured to be from about 5 to 30 ohms, for antenna heights from 1/16 to ¼ wavelength.

Top Loading

Because of the difficulty of obtaining a really low-resistance ground system, it is always desirable to make a grounded vertical antenna as high as possible, since this increases the radiation resistance. (There is no point in going beyond a half wavelength, however.) At the low frequencies where a grounded antenna is generally used, the heights required for the realization of high radiation resistance usually are impracticable for amateur work. The object of design of vertical grounded antennas which are necessarily ¼ wavelength or less high is to make the current loop come near the top of the antenna, and to keep the current as large as possible throughout the length of the vertical wire. This requires "top loading," which means replacing the missing height by some form of electrical circuit having the same characteristics as the missing part of the antenna, so far as energy traveling up to the end of the antenna is concerned.

One method of top loading is shown in Fig. 2-81. The vertical section of the antenna terminates in a "flat-top" which supplies a capacitance at the top into which current can flow. The simple single-conductor system shown at A is more readily visualized as a continuation of the antenna so that the dimension "X" is essentially the over-all length of the antenna. If this dimension is a half wavelength, the resistance at the antenna terminals (indicated by the small circles, one being grounded) will be high. A

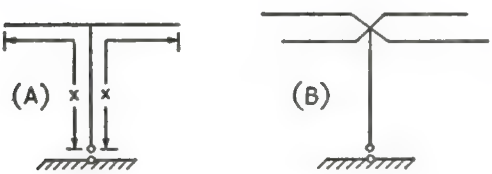


Fig. 2-81 — Simple top loading of a vertical antenna. The antenna terminals, indicated by the small circles, are the base of the antenna and ground, and should not be taken to include the length of any lead-ins or connecting wires.

disadvantage of this system is that the horizontal portion also radiates to some extent, although there is cancelation of radiation in the direction at right angles to the wire direction, since the currents in the two portions are flowing in opposite directions.

A multiwire system such as is shown at B will have more capacitance than the single-conductor arrangement and thus will not need to be as long, for resonating at a given frequency, but requires extra supports for the additional wires. Ideally, an arrangement of this sort should be in the form of a cross, but parallel wires separated by several feet will give a considerable increase in capacitance over a single wire.

With either system dimension "X," the length from the base of the antenna along one conductor to the end, should be not more than one-half wavelength nor less than one-fourth wavelength.

Instead of a flat top, it is possible to use a simple vertical wire with concentrated capacitance and inductance at its top to simulate the effect of the missing length. The capacitance used is not the usual type of capacitor, which would be ineffective since the connection is one-

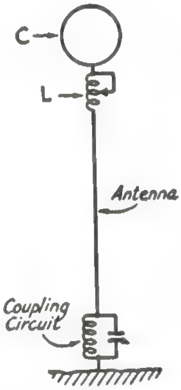


Fig. 2-82 — Top loading with lumped constants. The inductance, *L, should* be adjusted to give maximum field strength with constant power input to the antenna. A parallel-tuned circuit, independently resonant at the operating frequency, is required for coupling to the transmitter when the top loading is adjusted to bring a current node at the lower end of the antenna.

sided, but consists of a metallic structure large enough to have the necessary self capacitance. Practically any sufficiently-large metallic structure can be used for the purpose, but simple geometric forms such as the sphere, cylinder and disk are preferred because of the relative ease with which their capacitance can be calculated. The inductance may be the usual type of r.f. coil, with suitable protection from the weather.

The minimum value of capactive reactance required depends principally upon the ground resistance. Fig. 2-83 is a set of curves giving the reactances required under representative conditions. These curves are based on obtaining 75 per cent of the maximum possible increase in field strength over an antenna of the same height without top loading, and apply with sufficient accuracy to all antenna heights. An inductance coil of reasonably low-loss construction is assumed. The general rule is to use as large a capacitance (low capacitive reactance) as the circumstances will permit, since an increase in capacitance will cause an improvement in the field strength. It is particularly important to do this when, as is usually the case, the ground resistance is not known and cannot be measured.

The capacitance of three geometric forms is

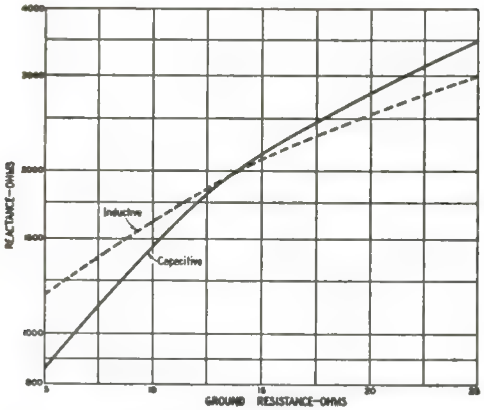


Fig. 2-83 — Inductive and capacitive reactance required for top loading a grounded antenna by the method shown in Fig. 2-81. The reactance values should be converted to inductance and capacitance, using the ordinary formulas, at the operating frequency.

shown by the curves of Fig. 2-84 as a function of their size. For the cylinder, the length is specified equal to the diameter. The sphere, disk and cylinder can be constructed from sheet metal, if such construction is feasible, but the capacitance will be practically the same in each case if a "skeleton" type of construction, using screening or wire networks, is used.

Ground-Plane Antennas

Instead of being actually grounded, a ¼-wave antenna can work against a simulated ground called a **ground plane.** Such a simulated ground can be formed from wires ¼ wavelength long radiating from the base of the antenna as shown in Fig. 2-85. It is obvious that with ¼-wave radials the antenna and any one radial have a total length of ½ wavelength and therefore will be a resonant system. However, with only one radial the directive pattern would be that of a half-wave antenna bent into a right angle at the center; if one section is vertical and the other horizontal this would result in equal components of horizontal and vertical polarization and a nonuniform pattern in the horizontal plane. This can be overcome by using a ground plane in the shape of a disk with a radius of ¼ wavelength. The effect of the disk can be simulated, with simpler construction, by using at least four straight radials equally spaced around the circle, as indicated in the drawing.

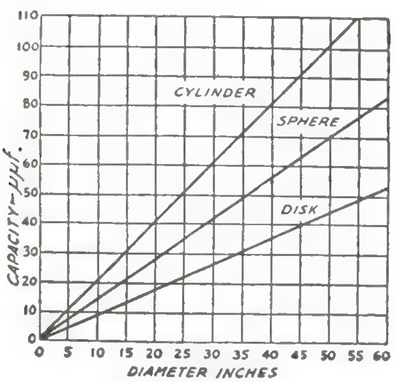


Fig. 2-84 — Capacitance of sphere, disk and cylinder as a function of their diameters. The cylinder length is assumed equal to its diameter.

The ground-plane antenna is widely used at v.h.f., for the purpose of establishing a "ground" for a vertical antenna mounted many wavelengths above actual ground. This prevents a metallic antenna support from carrying currents that tend to turn the system into the equivalent of a vertical long-wire antenna and thus raising the wave angle.

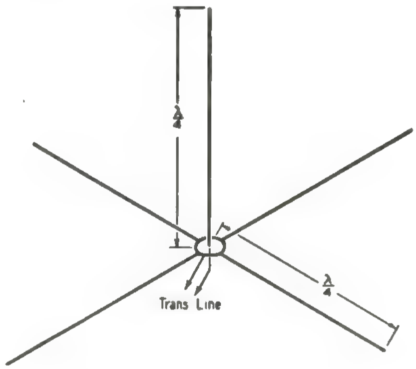


Fig. 2-85 — The ground-plane antenna. Power is applied between the base of the antenna and the center of the ground plane, as indicated in the drawing.

At frequencies of the order of 14 to 30 Mc. a ground plane of the type shown in Fig. 2-85 will permit using a quarter-wave vertical antenna (for nondirectional low-angle operation) a a height that will let the antenna be clear of its surroundings. Such short antennas mounted on the ground itself are frequently so surrounded by energy-absorbing structures and trees as to be rather ineffective. Since the quarter-wave radials are physically short at these frequencies, it is quite practical to mount the entire system on a roof top or pole. A ground plane at a height of a halfwave length or more closely approximates perfectly conducting earth, and the resistance curve of Fig. 2-78 applies with good accuracy. The antenna itself may be any desirable height and does not have to be exactly a quarter wavelength. The radials, however, preferably should be quite close to a quarter wave in length.

A ground plane also can be of considerable benefit at still lower frequencies provided the radials and the base of the antenna can be a moderate height above the ground. In order to take over the function of the actual ground connection, the ground plane must be so disposed that the field of the antenna prefers to travel

along the ground-plane wires instead of in the ground itself, thus confining the current to the highly-conducting wires rather than letting it flow in lossy earth. This is particularly necessary where the current is greatest; i.e., close to the antenna. If the ground plane has to be near the earth, the number of wires should be increased, using as many as is practicable and spacing them as evenly as possible in a circle around the antenna. If the construction of a multiwire ground plane is impracticable, a better plan is to bury as many radials as possible in the ground as described earlier.

SHORT ANTENNAS IN GENERAL

Although the discussions in this chapter have principally been in terms of self-resonant antennas, particularly those a half-wavelength long, it would be a mistake to assume that there is anything peculiarly sacred about resonance. The resonant length happens to be one that it is convenient to analyze. At other lengths the directive properties will be different, the radiation resistance will be different, and the impedance looking into the terminals of the antenna will contain reactance as well as resistance. The reactance presents no problem, because it can easily be "tuned out" and the antenna *system* thereby resonated even though the antenna by itself is *not* resonant. The purpose of resonating either the antenna or the system as a whole is simply to facilitate feeding power to the antenna, and such resonating or tuning does not affect the antenna's radiating properties.

Physical conditions frequently make it necessary to use antennas shorter than a half wavelength. The directive pattern of a short antenna does not differ greatly from that of a half-wave antenna, and in the limit the pattern approaches that shown in Fig. 2-12. The difference in the field strength that this shift in pattern shape causes is negligible. The most important difference is the decrease in radiation resistance and its effect on the efficiency of the antenna. This has been discussed in the preceding section on grounded antennas. The curve of Fig. 2-78 can be used for any center-fed nongrounded antenna by using half the actual antenna length and multiplying the corresponding radiation resistance by two. For example, an antenna having an actual length of 120 degrees (⅓ wavelength) has a half length of 60 degrees and a radiation resistance of about 13 ohms per side or 26 ohms for the whole antenna. The reactance, which will be capacitive and of the order of several hundred ohms, can be tuned out by a loading coil or coils. As described earlier, low-resistance coils must be used if the antenna efficiency is to be kept reasonably high. However, the ground resistance loss can be neglected in a horizontal center-fed antenna of this type if the height is a quarter wavelength or more.

A short antenna should not be made shorter than the physical circumstances require, because the efficiency decreases rapidly as the antenna is made shorter. For example, a center-fed antenna having an over-all length of ¼ wavelength (half length 45 degrees) will have a radiation resistance of $2 \times 7 = 14$ ohms, as shown by Fig. 2-79. Depending on the length/diameter ratio, the reactance will be 500 to 1000 ohms. A loading coil of the same reactance probably will have a resistance of 3 to 6 ohms, so the probable efficiency will be 70 to 80 per cent, or a loss of 1 to 1.5 db. While this is not too bad, further shortening not only further decreases the radiation resistance but enters a length region where the reactance increases very rapidly, so that the coil resistance quickly becomes larger than the radiation resistance. Where the antenna must be short, a small length/diameter ratio (thick antenna) is definitely desirable as a means of keeping down the reactance and thus reducing the size of loading inductance required.

LOOP ANTENNAS

A loop antenna is a closed-circuit antenna—that is, one in which a conductor is formed into one or more turns so that its two ends are close together. Loops can be divided into two general classes, those in which both the total conductor length and the maximum linear dimension of a turn are both very small compared with the wavelength, and those in which both the conductor length and the loop dimensions begin to be comparable with the wavelength.

A "small" loop can be considered to be simply a rather large coil, and the current distribution in such a loop is the same as in a coil. That is, the current is in the same phase and has the same

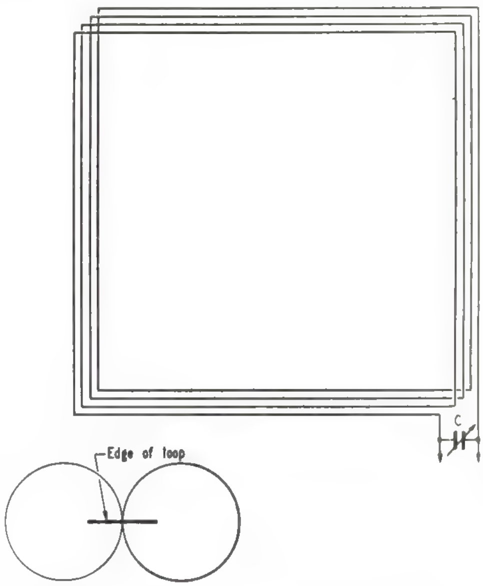


Fig. 2-86 — Small loop, consisting of several turns of wire having a total length very much less than a wavelength. The directional pattern of such a loop is as shown in the small drawing, with maximum response in the plane of the loop.

amplitude in every part of the loop. To meet this condition the total length of conductor in the loop must not exceed about 0.08 wavelength.

A "large" loop is one in which the current is not the same either in amplitude or phase in every part of the loop. This change in current distribution gives rise to entirely different properties as compared with a small loop.

Small Loops

Small loops can be made in the form of a circle, triangle, rectangle, etc., with little or no change in properties. The most convenient form, generally, is a square such as is shown in Fig. 2-86. So long as the total length of the conductor is very small compared with the wavelength the loop acts like a simple inductance and can be turned to resonance at the desired frequency by a condenser, *C*. The directive pattern of such a loop is given by the small drawing, and is the same as that of an elementary doublet (Fig. 2-12).

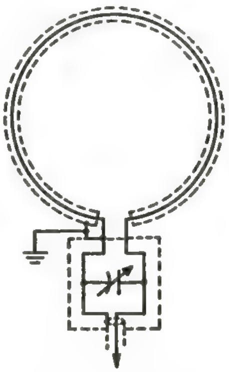


Fig. 2-87 — Shielded loop for direction finding. The ends of the shielding turn are insulated from each other to prevent shielding the loop from magnetic fields, although the shielding is effective against electric fields.

Loops of this type do not have much application in amateur work, although they are widely used at frequencies below the standard broadcast band for direction finding. They are not very useful for this purpose at high frequencies because waves arriving at a receiving point via the ionosphere have random polarization and wave angles, and this introduces large errors. However, they are capable of giving good results for this purpose if only the ground wave is present, provided the location of the loop is such that false bearings are not caused by reflections from near-by conductors. Shielded loops have been used with considerable success in "transmitter hunts" utilizing frequencies in the 28-Mc. band. The shield, shown schematically in Fig. 2-87, is used for preventing "antenna effect"—that is, to eliminate undesired response of the loop considered merely as a mass of metal connected to the receiver antenna terminals.

The radiation resistance of a small loop is extremely low. For this reason most of the power supplied to the loop is wasted in conductor resistance loss, when the loop is used for transmitting. A similar situation exists when the loop is

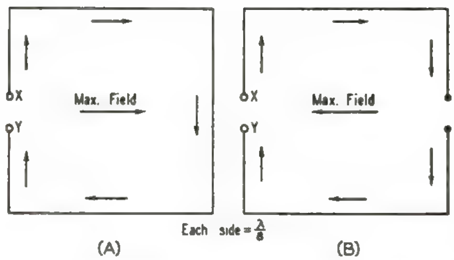


Fig. 2-88 — Half-wave loops, consisting of a single turn having a total length of ½ wavelength.

used for receiving; because of its small size only a relatively small amount of energy is absorbed from passing waves. However, at comparatively low frequencies such as the 3.5-Mc. band it can draw energy from a fairly large area (see earlier section on pick-up efficiency) of the passing wave front and as a result may establish a reasonable signal-to-noise ratio at the receiver's input terminals. It should be tuned to the received frequency in such case, and when so resonated may be markedly better than the few feet of wire often used for reception on the lower frequencies. Also, it is sometimes possible to take advantage of its directional effects to reduce interference.

Half-Wave Loops

The smallest size of "large" loop generally used is one having a conductor length of ½ wavelength. The conductor is generally formed into a square, as shown in Fig. 2-88, making each side ⅛ wavelength long. When fed at the center of one side the current flows in a closed loop as shown at A. The current distribution is approximately the same as on a half-wave wire, and so is maximum at the center of the side opposite the terminals *X-Y*, and minimum at the terminals themselves. This current distribution causes the field strength to be maximum in the plane of the loop and in the direction looking from the low-current side to the high-current side. If the side opposite the terminals is opened at the center as shown at B (strictly speaking, it is then no longer a loop because it is no longer a closed circuit) the direction of current flow remains unchanged but the maximum current occurs at the terminals. This reverses the direction of maximum radiation.

The radiation resistance at a current antinode (which is also the resistance at *X-Y* in Fig. 2-88B) is of the order of 50 ohms. The impedance at the terminals in A is a few thousand ohms. This can be reduced by using two identical loops side by side with a few inches spacing between them and applying power between terminal *X* on one loop and terminal *Y* on the other.

Unlike a half-wave dipole or a small loop, there is no direction in which the radiation from a loop of the type shown in Fig. 2-88 is zero. There is appreciable radiation in the direction

perpendicular to the plane of the loop, as well as to the "rear"—the opposite direction to the arrows shown. The front-to-back ratio is of the order of 4 to 6 db. The small size and the shape of the directive pattern result in a loss of about 1 db. when the field strength in the optimum direction from such a loop is compared with the field from a half-wave dipole in its optimum direction.

The ratio of the forward radiation to the backward radiation can be increased and the field strength likewise increased at the same time to give a gain of about 1 db. over a dipole, by using inductive reactances to "load" the sides joining the front and back of the loop. This is shown in Fig. 2-89. The reactances, which should have a value of approximately 360 ohms, decrease the current in the sides in which they are inserted and increases it in the side having the terminals. This increases the directivity and thus increases the efficiency of the loop as a radiator.

One-Wavelength Loops

Loops in which the conductor length is one wavelength (sides of square equal to ¼ wavelength) have different characteristics than half-wave loops. Two forms of one-wavelength loops are shown in Fig. 2-90, the difference being in the point at which the terminals are inserted. The relative direction of current flow is as shown in the drawings. This direction reverses halfway around the perimeter of the loop, since such rehearsals always occur at the junction of each halfwave section of wire.

The directional characteristics of loops of this type are opposite in sense to those of a small loop. That is, the radiation is maximum perpendicular to the plane of the loop and is minimum in any direction in the plane containing the loop. If the two loops shown in Fig. 2-90 are mounted in a vertical plane with the terminals at the bottom, the radiation is horizontally polarized. When the terminals are moved to the center of one vertical side in A, or to a side corner in B, the radiation is vertically polarized.

In contrast to straight-wire antennas, the electrical length of the circumference of a one-wavelength loop is *shorter* than the actual length. For loops made of wire and operating at frequencies below 30 Mc. or so, where the ratio of conductor length to wire diameter is large, the loop will be close to resonance when

$$\text{Length in feet} = \frac{1005}{f\,\text{Mc.}}$$

The radiation resistance of a resonant one-wavelength loop is 100 ohms, approximately, when the ratio of conductor length to diameter is large. As the loop dimensions are comparable with those of a half-wave dipole, the radiation efficiency is high.

In the direction of maximum radiation (that is, broadside to the plane of the loop, regardless of the point at which it is fed) the one-wavelength loop will show a small gain over a half-wave dipole. Theoretically, this gain is about 2 db., and measurements have confirmed that it is of this order.

The one-wavelength loop is more frequently used as an element of a directive antenna array (the "quad antenna" described in a later chapter) than singly, although there is no reason why it cannot be used alone. In the quad, it is nearly always driven so that the polarization is horizontal.

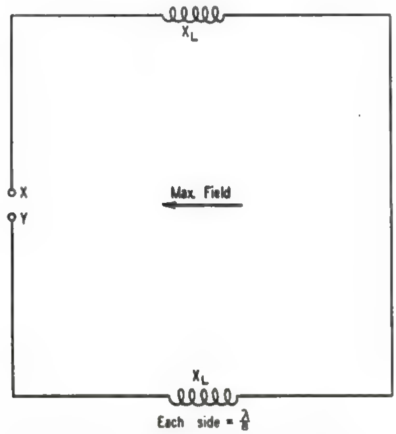


Fig. 2-89 — Inductive loading in the sides of a half-wave loop to increase the directivity and gain. Maximum radiation or response is in the plane of the loop in the direction shown by the arrow.

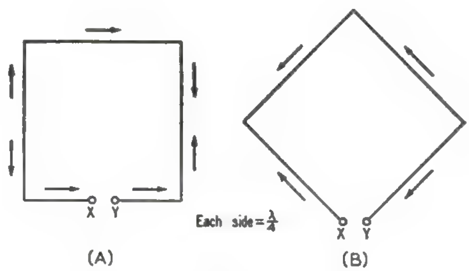


Fig. 2-90 — Loops having sides ¼ wavelength long (total conductor length 1 wavelength). The polarization depends on the orientation of the loop and the point at which the terminals X–Y are located.

OTHER TYPES OF ANTENNAS

The half-wave dipole and the few special types of antennas described in this chapter form the basis for practically all antenna systems in use at frequencies from the v.h.f. region down. Other fundamental types of radiators are applicable at microwaves, but they are not used at lower frequencies because the dimensions are such as to be wholly impracticable when the wavelength is measured in meters rather than centimeters.

Transmission Lines

The desirability of installing an antenna in a clear space, not too near buildings or power and telephone lines, has been emphasized in the preceding chapter. On the other hand, the transmitter that generates the r.f. power for driving the antenna is usually, as a matter of necessity, located at some distance from the antenna terminals. The connecting link between the two is the r.f. transmission line or feeder. Its sole purpose is to carry r.f. power from one place to another, and to do it as efficiently as possible. That is, the ratio of the power *transferred* by the line to the power *lost* in it should be as large as the circumstances will permit.

At radio frequencies every conductor that has appreciable length compared with the wavelength in use will *radiate* power. That is, every conductor becomes an antenna. Special care must be used, therefore, to minimize radiation from the conductors used in r.f. transmission lines. Without such care, the power radiated by the line may be much larger than that which is lost in the resistance of conductors and dielectrics. Power loss in resistance is inescapable, at least to a degree, but loss by radiation is largely avoidable.

Preventing Radiation

Radiation loss from transmission lines can be prevented by using two conductors so arranged and operated that the electromagnetic field from one is everywhere balanced by an equal and opposite field from the other. In such a case the resultant field is zero everywhere in space; in other words, there is no radiation.

For example, Fig. 3-1A shows two parallel conductors having currents I_1 and I_2 flowing in opposite directions. If the current I_1 at point Y on the upper conductor has the same amplitude as the current I_2 at the corresponding point X on the lower conductor, the fields set up by the two currents will be equal in magnitude. Because the two currents are flowing in opposite directions, the field from I_1 at Y will be 180 degrees out of phase with the field from I_2 at X. However, it takes a measurable interval of time for the field from X to travel to Y. If I_1 and I_2 are alternating currents, the phase of the field from I_1 at Y will have changed in such a time interval, and so at the instant the field from X reaches Y the two fields at Y are not exactly 180 degrees out of phase. The two fields will be exactly 180 degrees out of phase at every point in space only when the two conductors occupy the same space —an obviously impossible condition if they are to remain separate conductors.

The best that can be done is to make the two fields cancel each other as completely as possible. This can be accomplished by making the distance, d, between the two conductors small enough so that the time interval during which

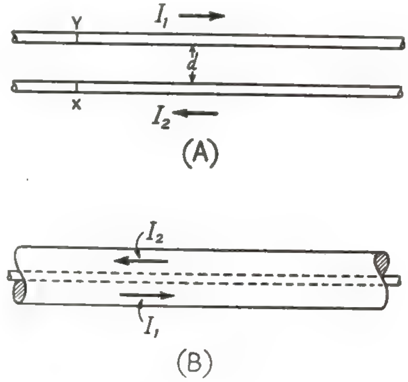


Fig. 3-1—The two basic types of transmission lines.

the field from X is moving to Y is a very small part of a cycle. When this is the case the phase difference between the two fields at any given point will be so close to 180 degrees that the cancellation is practically complete.

Practicable values of d, the separation between the two conductors, are determined by the physical limitations of line construction. A separation that meets the condition of being "very small" at one frequency may be quite large at another. For example, if d is six inches the phase difference between the two fields at Y will be only a fraction of a degree if the frequency is 3500 kc. This is because a distance of six inches is such a small fraction of a wavelength (one wavelength = 360 degrees) at 3500 kc. But at 144 Mc. the phase difference would be 26 degrees, and at 420 Mc. it would be 73 degrees. In neither of these cases could the two fields be considered to "cancel" each other. The separation must be very small in comparison with the wavelength used; it should never exceed 1 per cent of the wavelength and smaller separations are desirable.

Transmission lines consisting of two parallel conductors as in Fig. 3-1A are called parallel-conductor lines, or open-wire lines, or two-wire lines.

A second general type of line construction is shown in Fig. 3-1B. In this case one of the conductors is tube-shaped and encloses the other conductor. This is called a coaxial line ("coax")

or concentric line. The current flowing on the inner conductor is balanced by an equal current flowing in the opposite direction on the inside surface of the outer conductor. Because of skin effect the current on the inner surface of the tube does not penetrate far enough to appear on the outer surface. In fact, the total electromagnetic field outside the coaxial line, as a result of currents flowing on the conductors inside, always is zero because the tube acts as a shield at radio frequencies. The separation between the inner conductor and the outer conductor is therefore unimportant from the standpoint of reducing radiation.

CURRENT FLOW IN LONG LINES

In Fig. 3-2, imagine that the connection between the battery and the two wires is made instantaneously and then broken. During the time the wires are in contact with the battery terminals, electrons in wire No. 1 will be attracted toward the positive battery terminal and an equal number of electrons in wire No. 2 will be repelled away from the negative terminal. This happens only near the battery terminals at first, since electrical effects do not travel at infinite speed, so some time will elapse before the currents become evident at more extreme parts of the wires. By ordinary standards the elapsed time is very short, since the speed of travel along the wires may be almost 300,000,000 meters per second, so it becomes necessary to measure time in millionths of a second (microseconds) rather than in more familiar time units.

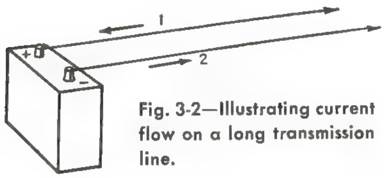


Fig. 3-2—Illustrating current flow on a long transmission line.

For example, suppose that the contact with the battery is so short that it can be measured in a very small fraction of a microsecond. Then the "pulse" of current that flows at the battery terminals during this time can be represented by the vertical line in Fig. 3-3. At the speed of light, this pulse will travel 30 meters along the line in 0.1 microsecond; 30 meters more, making a total of 60 meters, in 0.2 microsecond; a total of 90 meters in 0.3 microsecond, and so on for as far as the line reaches. The current does not exist all along the wires but is only present at the point that the pulse has reached in its travel; at this point it is present in both wires, with the electrons moving in one direction in one wire and in the other direction in the other wire. If the line is infinitely long and has no resistance or other cause of energy loss, the pulse will travel undiminished forever.

Extending the example of Fig. 3-3, it is not hard to see that if instead of one pulse we started a whole series of them on the line at equal time intervals, they would travel along the line with the same time and distance spacing between them, each pulse independent of the others. In fact, each pulse could have a different amplitude, if the battery voltage were varied to that end. Furthermore, the pulses could be so closely spaced that they touched each other, in which case we should have current present everywhere along the line simultaneously.

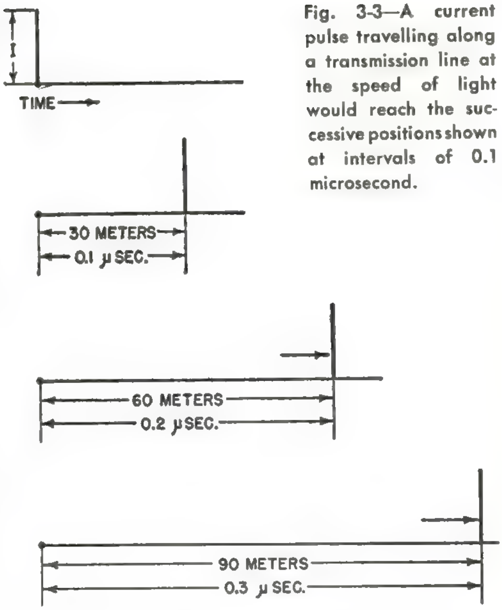


Fig. 3-3—A current pulse travelling along a transmission line at the speed of light would reach the successive positions shown at intervals of 0.1 microsecond.

Wavelength

It follows from this that an alternating voltage applied to the line would give rise to the sort of current flow shown in Fig. 3-4. If the frequency of the a.c. voltage is 10,000,000 cycles per second (10 Mc.) each cycle will occupy 0.1 microsecond, so a complete cycle of current will be present along each 30 meters of line. This is a distance of one wavelength. Any currents observed at *B* and *D* occur just one cycle later in time than the currents at *A* and *C*. To put it another way, the currents initiated at *A* and *C* do not appear at *B* and *D*, one wavelength away, until the applied voltage has had time to go through a complete cycle.

Since the applied voltage is always changing, the currents at *A* and *C* are changing in proportion. The current a short distance away from *A* and *C*—for instance, at *X* and *Y*—is not the same as the current at *A* and *C* because the current at *X* and *Y* was caused by a value of voltage that occurred slightly earlier in the cycle. This is true all along the line; at any instant the current anywhere along the line from *A* to *B* and *C* to *D* is different from the current at every other point in that same distance. The series of drawings shows how the instantaneous currents might be distributed if we could snapshot them at intervals of one-quarter cycle. The current travels out

from the input end of the line in waves. At any selected point on the line the current goes through its complete range of a.c. values in the time of one cycle just as it does at the input end. Therefore (if there are no losses) an ammeter inserted in either conductor would read exactly the same current at any point along the line, because the ammeter averages the current over a whole cycle. The *phases* of the currents at any two separated points would be different, but an ammeter cannot show phase.

Velocity of Propagation

In the example above it was assumed that energy travelled along the line with the velocity of light. The actual velocity is very close to that of light in a line in which the insulation between conductors is solely air. The presence of dielectrics other than air reduces the velocity, since electro-magnetic waves travel more slowly in dielectrics than they do in a vacuum. Because of this the wavelength as measured along the line will depend on the velocity factor that applies in the case of the particular type of line in use. (See later section in this chapter for actual figures.) The wavelength in a practical line is always shorter than the wavelength in free space.

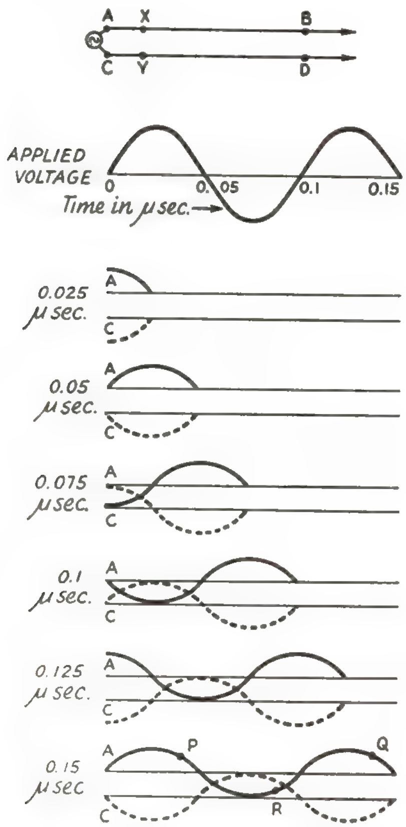


Fig. 3-4—Instantaneous current along a transmission line at successive time intervals. The frequency is such that the time of one cycle is 0.1 microsecond.

CHARACTERISTIC IMPEDANCE

If this is a "perfect" line—one without resistance—a question immediately comes up: what is the amplitude of the current in the pulse? Will a larger voltage result in a larger current, or is the current theoretically infinite for any applied voltage, as we would expect from applying Ohm's Law to a circuit without resistance? The answer is that the current does depend directly on the voltage, just as though resistance were present.

The reason for this is that the current flowing in the line is something like the charging current that flows when a battery is connected to a capacitor. That is, the line has capacitance. However, it also has inductance. Both of these are "distributed" properties. We may think of the line as being composed of a whole series of small inductors and capacitors connected as in Fig. 3-5, where each coil is the inductance of an ex-

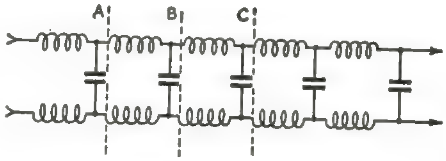


Fig. 3-5—Equivalent of a transmission line in terms of ordinary circuit constants. The values of *L* and *C* depend on the line construction.

tremely small section of wire and the capacitance is that existing between the same two sections. Each inductance limits the rate at which each immediately-following capacitor can be charged, and the effect of the chain is to establish a definite relationship between current and voltage. Thus the line has an apparent "resistance," called its characteristic resistance—or, a more general term, its characteristic impedance or surge impedance. The conventional symbol for characteristic impedance is Z_0.

TERMINATED LINES

The value of the characteristic impedance is equal to $\sqrt{L/C}$ in a perfect line—i.e., one in which the conductors have no resistance and there is no leakage between them—where L and C are the inductance and capacitance, respectively, per unit length of line. The inductance decreases with increasing conductor diameter, and the capacitance decreases with increasing spacing between the conductors. Hence a line with large conductors closely spaced will have relatively low characteristic impedance, while one with thin conductors widely spaced will have high impedance. Practicable values of Z_0 for parallel-conductor lines range from about 200 to 800 ohms and for typical coaxial lines from 50 to 100 ohms.

Matched Lines

In this picture of current travelling along a transmission line we have assumed that the line

was infinitely long. Practical lines have a definite length, and they are connected to or terminated in a load at the "output" end, or end to which the power is delivered. If the load is a pure resistance of a value equal to the characteristic impedance of the line, Fig. 3-6, the current travelling along the line to the load does not

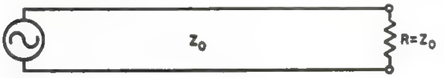


Fig. 3-6—A transmission line terminated in a resistive load equal to the characteristic impedance of the line.

find conditions changed in the least when it meets the load; in fact, the load just "looks like" still more transmission line of the same characteristic impedance.

The reason for this can perhaps be made a little clearer by considering it from another viewpoint. In flowing along a transmission line, the power is handed from one of the elementary sections in Fig. 3-5 to the next. When the line is infinitely long this power transfer always goes on in one direction—away from the source of power. From the standpoint of section *B*, Fig. 3-5, for instance, the power it has handed over to section *C* has simply disappeared in *C*. So far as section *B* is concerned, it makes no difference whether *C* has absorbed the power itself or has in turn handed it along to more line. Consequently, if we substitute something for section *C* that has the same electrical characteristics, section *B* will not know the difference. A pure resistance equal to the characteristic impedance of *C*, which is also the characteristic impedance of the line, meets this condition. It absorbs all the power just as the infinitely-long line absorbs all the power transferred by section *B*.

A line terminated in a purely-resistive load equal to the characteristic impedance is said to be matched. In a matched transmission line, power travels outward along the line from the source until it reaches the load, where it is completely absorbed. Thus with either the infinitely-long line or its matched counterpart the impedance presented to the source of power (the line input impedance) is the same regardless of the line length. It is simply equal to the characteristic impedance of the line. The current in such a line is equal to the applied voltage divided by the characteristic impedance, and the power put into it is E^2/Z_0 or I^2Z_0, by Ohm's Law.

Mismatched Lines

Now take the case where the terminating resistance, *R*, is *not* equal to Z_0, as in Fig. 3-7. The load *R* no longer "looks like" more line to the section of line immediately adjacent. Such a line is said to be mismatched. The more *R* differs from Z_0 the greater the mismatch. The power reaching *R* is not totally absorbed, as it was when *R* was equal to Z_0, because *R* requires a different voltage-to-current ratio than the one at which the power is travelling along the line. The result is that *R* absorbs only part of the power reaching it (the **incident** power); the remainder acts as though it had bounced off a wall and starts back along the line toward the source. This is **reflected power** and the greater the mismatch the larger the percentage of the incident power that is reflected. In the extreme case where *R* is zero (a short circuit) or infinity (an open circuit) *all* of the power reaching the end of the line is reflected.

Whenever there is a mismatch power is travelling in both directions along the line. The voltage-to-current ratio is the same for the reflected power as for the incident power, since this ratio is determined by the Z_0 of the line. The voltage and current travel along the line in both directions in the same sort of wave motion shown in Fig. 3-4. When the source of power is an a.c. generator, the outgoing or incident voltage and the returning or reflected voltage are simultaneously present all along the line, so the actual voltage at any point along the line is the sum of the two components, taking phase into account. The same thing is true of the current.

The effect of the incident and reflected components on the behavior of the line can be understood more readily by considering first the two limiting cases—the short-circuited line and the open-circuited line. If the line is short-circuited as in Fig. 3-7B, the voltage at the end must be zero. Thus the incident voltage must disappear suddenly at the short. It can do this only if the reflected voltage is opposite in phase and of the same amplitude. This is shown by the vectors in Fig. 3-8. The current, however, does not disappear in the short circuit; in fact, the incident current flows through the short and there is in addition the reflected component in phase with it and of the same amplitude. The reflected voltage and current must have the same amplitudes as the incident voltage and current because no power is used up in the short circuit; all the power starts back toward the source. Reversing the phase of *either* the current or voltage (but not both) will reverse the direction of power flow; in the short-circuited case the phase of the voltage is reversed on reflection but the phase of the current is not.

If the line is open-circuited (Fig. 3-7C) the

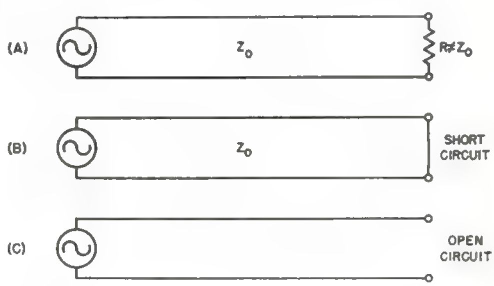


Fig. 3-7—Mismatched lines. A—termination not equal to Z_0; B—short-circuited line; C—open-circuited line.

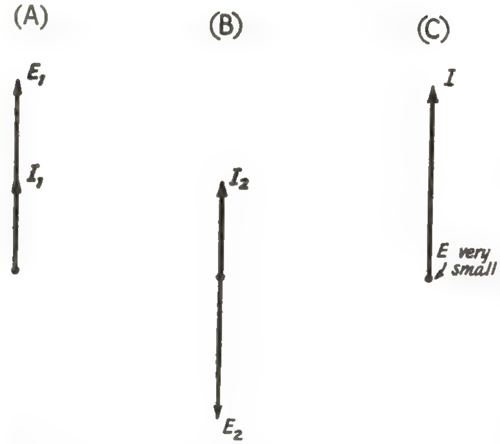


Fig. 3-8—Voltage and current at the short-circuit on a short-circuited line. These vectors show how the outgoing voltage and current (A) combine with the reflected voltage and current (B) to result in high current and very low voltage in the short-circuit.

current must be zero at the end of the line. In this case the reflected current is 180 degrees out of phase with the incident current and has the same amplitude. By reasoning similar to that used in the short-circuited case, the reflected voltage must be in phase with the incident voltage, and must have the same amplitude. Vectors for the open-circuited case are shown in Fig. 3-9.

Where there is a finite value of resistance at the end of the line, Fig. 3-7A, only part of the power reaching the end of the line is reflected. That is, the reflected voltage and current are smaller than the incident voltage and current. If R is less than Z_0 the reflected and incident voltages are 180 degrees out of phase, just as in the case of the short-circuited line, but the amplitudes are not equal because all of the voltage does not disappear at R. Similarly, if R is greater than Z_0 the reflected and incident currents are 180 degrees out of phase, as they were in the open-circuited line, but all of the current does not disappear in R so the amplitudes of the two components are not equal. These two cases are shown in Fig. 3-10. Note that the resultant current and voltage are in phase in R, since R is a pure resistance.

Reflection Coefficient

The ratio of the reflected voltage to the incident voltage is called the **reflection coefficient.** Thus

$$k = \frac{E_r}{E_f}$$

where k is the reflection coefficient, E_r is the reflection voltage, and E_f is the incident or forward voltage. The reflection coefficient is determined by the relationship between the line Z_0 and the actual load at the terminated end of the line. For any given line and load it is a constant if the line has negligible loss in itself. The coefficient can never be larger than 1 (which indicates that all the incident power is reflected) nor smaller than zero (indicating that the line is perfectly matched by the load).

If the load is purely resistive, the reflection coefficient can be found from

$$k = \frac{R - Z_0}{R + Z_0}$$

where R is the resistance of the load terminating the line. In this expression k is positive if R is larger than Z_0 and negative if R is smaller than Z_0. The change in signs accompanies the change in phase of the reflected voltage described above.

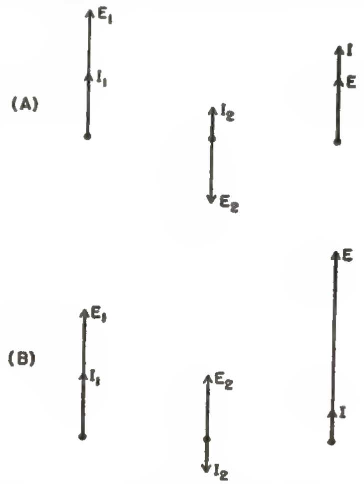


Fig. 3-10—Incident and reflected components of voltage and current when the line is terminated in a pure resistance not equal to Z_0. In the case shown, the reflected components have half the amplitude of the incident components. A—R less than Z_0; B—R greater than Z_0.

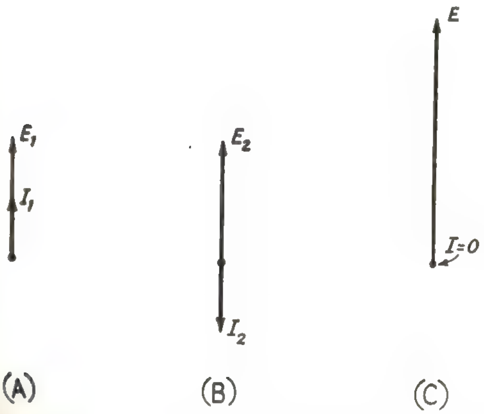


Fig. 3-9—Voltage and current at the end of an open-circuited line. A—outgoing voltage and current; B—reflected voltage and current; C—resultant.

STANDING WAVES

As might be expected, reflection cannot occur at the load without some effect on the voltages and currents all along the line. A detailed description tends to become complicated, and what happens is most simply shown by vector diagrams. Fig. 3-11 is an example in the case where R is less than Z_0. The voltage and current age is everywhere the same (and similarly for the current). The mismatch between load and line is responsible for the variations in amplitude which, because of their wavelike appearance are called standing waves.

From the earlier discussion it should be clear that when R is greater than Z_0, the voltage will be largest and the current smallest at the load. This is just the reverse of the case shown in Fig.

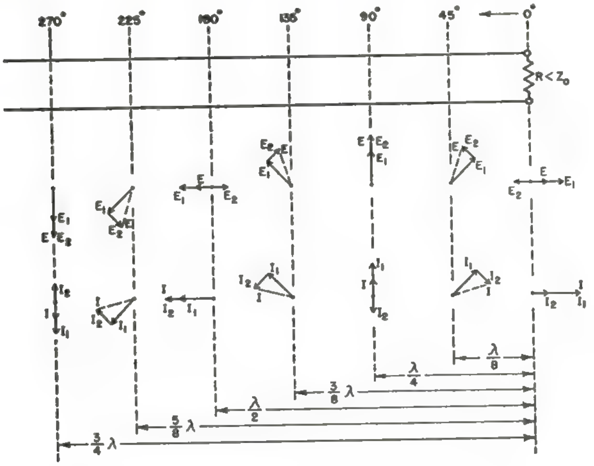


Fig. 3-11—Incident and reflected components at various positions along the line, together with resultant voltages and currents at the same positions. The case shown is for R less Z_0.

vectors at the load R, are shown in the reference position; they correspond with the vectors in Fig. 3-10A. Going back along the line from R towards the power source, the incident vectors, E_1 and I_1, lead the vectors at the load according to their position along the line measured in electrical degrees. (The corresponding distances in fractions of a wavelength also are shown.) The vectors representing reflected voltage and current, E_2 and I_2, successively lag the same vectors at the load. This lag and lead is the natural consequence of the direction in which the incident and reflected components are travelling, together with the fact that it takes time for the power to travel along the line. The resultant voltage, E, and current, I, at each of these positions are shown dotted. Although the incident and reflected components maintain their respective amplitudes (the reflected component is shown at half the incident-component amplitude in this drawing) their phase relationships vary with position along the line. The phase shifts cause both the amplitude and phase of the *resultants* to vary with position on the line.

If the amplitude variations (disregarding phase) of the resultant voltage and current are plotted against position along the line, graphs like those of Fig. 3-12 will result. If we could go along the line with a voltmeter and ammeter plotting the current and voltage at each point, we should find that the data collected gave curves like these. In contrast, if the load matched the Z_0 of the line, similar measurements along the line would show that the volt-

3-12. In such case the curve labelled E would become the I (current) curve, while the current curve would become the voltage curve.

Some general conclusions can be drawn from inspection of the standing-wave curves: At a position 180 degrees (½ wavelength) from the load the voltage and current have the same values they do at the load. At a position 90 degrees from the load the voltage and current are "inverted." That is, if the voltage is lowest and the current highest at the load (R less than Z_0), then 90 degrees or ¼ wavelength from the load the voltage reaches its highest value and the current reaches its lowest value. In the case where R is greater than Z_0, so that the voltage is highest and the current lowest at the load, the voltage has its lowest value and the current its highest value at a point 90 degrees from the load.

Note that the conditions existing at the 90-

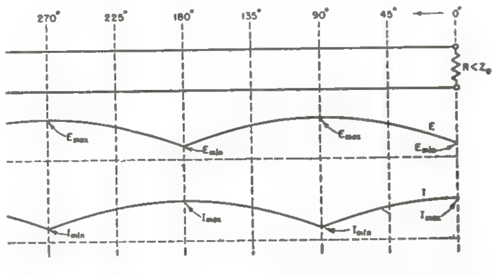


Fig. 3-12—Standing waves of current and voltage along the line, for R less than Z_0.

degree point also are duplicated at the 270-degree point (¾ wavelength). If the graph were continued on toward the source of power it would be found that this duplication occurs at every point that is an odd multiple of 90 degrees (odd multiple of a quarter wavelength) from the load. Similarly, the voltage and current are the same at every point that is a multiple of 180 degrees (any multiple of one-half wavelength) as they are at the load.

Standing-Wave Ratio

The ratio of the maximum voltage along the line to the minimum voltage—that is, the ratio of E_{max} to E_{min} in Fig. 3-12—is called the **voltage standing-wave ratio** (abbreviated **v.s.w.r.**) or simply the **standing-wave ratio (s.w.r.)**. The ratio of the maximum current to the minimum current (I_{max}/I_{min}) is the same as the v.s.w.r., so either current or voltage can be measured to determine the standing-wave ratio.

The standing-wave ratio is an index of many of the properties of a mismatched line. It can be measured with good accuracy with fairly simple equipment, and so is a convenient quantity to use in making calculations on line performance. It is numerically equal to the ratio between the load resistance, R, and the characteristic impedance of the line; that is,

$$S.W.R. = \frac{R}{Z_0}$$

when R is greater than Z_0, and

$$S.W.R. = \frac{Z_0}{R}$$

when R is less than Z_0. The smaller quantity is always used in the denominator of the fraction so the s.w.r. will be a number larger than 1.

This relationship shows that the greater the mismatch—that is, the greater the difference between Z_0 and R—the larger the s.w.r. In the case of open- and short-circuited lines the s.w.r. becomes infinite. On such lines the voltage and current become zero at the minimum points (E_{min} and I_{max}) since total reflection occurs at the end of the line and the incident and reflected components have equal amplitudes.

INPUT IMPEDANCE

The relationship between the voltage and current at any point along the line (including the effects of both the incident and reflected components) becomes more clear when only the resultant voltage and current are shown, as in Fig. 3-13. Note that the voltage and current are in phase not only at the load but also at the 90-degree point, the 180-degree point, and the 270-degree point. This is also true at every point that is a multiple of 90 degrees from the load.

Suppose the line were cut at one of these points and the generator or source of power were connected to the remaining portion—that terminated in R. Then the generator would "see" a pure resistance, just as it would if it were connected directly to R. However, the value of the resistance it sees would depend on the line length. If the length is 90 degrees, or an odd multiple of 90 degrees, where the voltage is high and the current low, the resistance seen by the generator would be greater than R. If the length is 180 degrees or a multiple of 180 degrees, the voltage and current relationships are the same as in R, and therefore the generator "sees" a resistance equal to the actual load resistance at these line lengths.

The current and voltage are exactly in phase only at points that are multiples of 90 degrees

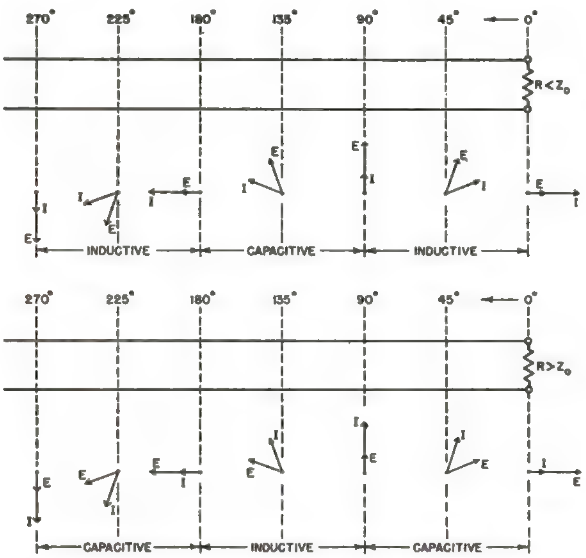


Fig. 3-13—Resultant voltages and currents along a mismatched line. Above—R less than Z_0; below—R greater than Z_0.

from the load. At all other points the current either leads or lags the voltage, and so the load seen by the generator when the line length is not an exact multiple of 90 degrees is not a pure resistance. The input impedance of the line—that is, the impedance seen by the generator connected to the line—in such a case has both resistive and reactive components. When the current lags behind the voltage the reactance is inductive; when it leads the voltage the reactance is capacitive. The upper drawing in Fig.

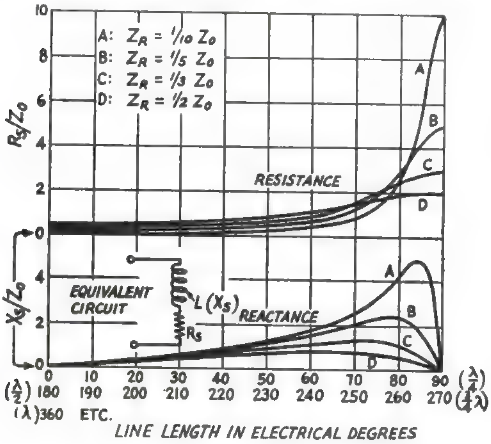


Fig. 3-14.—Universal curves of resistance and reactance vs. line length, for various Z_R/Z_0 ratios with Z_R less than Z_0. Actual values of resistance and reactance are found by multiplying the quantity R_S/Z_0 or X_S/Z_0 by the characteristic impedance of the line.

3-13 shows that when the line is terminated in a resistance smaller than Z_0 the reactance is inductive in the first 90 degrees of line moving from the load toward the generator, is capacitive in the second 90 degrees, inductive in the third 90 degrees, and so on every 90 degrees or quarter wave length. The lower drawing illustrates the case where *R* is greater than Z_0. The voltage and current vectors are merely interchanged, since, as explained in connection with Fig. 3-11, in this case the vector for the reflected current is the one that is reversed in phase on reflection. The reactance becomes capacitive in the first 90 degrees, inductive in the second, and so on.

Factors Determining the Input Impedance

The magnitude and phase angle of the input impedance depend on the s.w.r., the line length, and the Z_0 of the line. If the s.w.r. is small, the input impedance is principally resistive at all line lengths; if the s.w.r. is high, the reactive component may be relatively large. This is shown by the curves of Figs. 3-14 and 3-15 for standing-wave ratios of 10 (A), 5 (B), 3 (C) and 2 (D) to 1. The input impedance of the line can be represented by a series circuit of resistance and reactance, as shown by the inset drawings, where R_s is the resistive component and X_s is the reactive component. The values on the graphs are "normalized" to the Z_0 of the line; that is, the resistance and reactance are divided by Z_0 so the curves apply to lines of all characteristic impedances. For any given Z_0, actual values of resistance and reactance are found by multiplying the scaled value by Z_0. Thus in Fig. 3-14, curve A for a line 70 degrees long has a resistance value of about 0.8 and a reactance value of about 2.6. If the Z_0 of the line is 50 ohms, the equivalent circuit of the input impedance would consist of a resistance of $0.8 \times 50 = 40$ ohms in series with an inductive reactance of $2.6 \times 50 = 130$ ohms.

Equivalent Circuits for the Input Impedance

The series circuits shown in Figs. 3-14 and 3-15 are equivalent to the actual input imped-

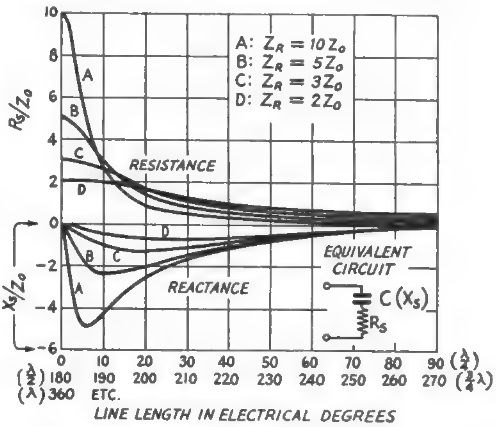


Fig. 3-15—Universal resistance-reactance curves for Z_R greater than Z_0. These curves are used in the same way as those of Fig. 3-14.

ance of the line because they have the same total impedance and the same phase angle. It is also possible to form a circuit with resistance and reactance in parallel that will have the same total impedance and phase angle as the line. This equivalence is shown in Fig. 3-16. The individual values in the parallel circuit are not the same as those in the series circuit (although the over-all result is the same) but are related to the series-circuit values by the equations shown in the drawing.

Either of the two equivalent circuits may be used, depending on which happens to be more convenient for the particular purpose. These circuits are important from the standpoint of designing coupling networks so that the desired amount of power will be taken from the source. Their practical application is described later in this chapter.

REACTIVE TERMINATIONS

So far the only type of load considered has been a pure resistance. In general, the actual load usually will be fairly close to being a pure resistance, since most transmission lines used by

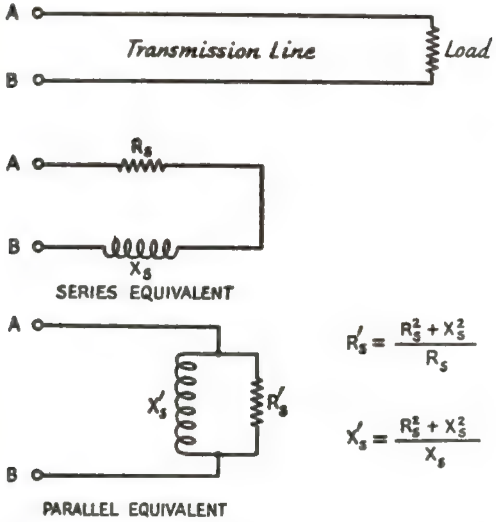


Fig. 3-16—Input impedance of a line terminated in a pure resistance. The input impedance can be represented either by a resistance and reactance in series or a resistance and reactance in parallel. The relationship between the series *R* and *X* values in the series and parallel equivalents is given by the formulas. *X* may be either inductive or capacitive, depending on the line length and the ratio Z_R/Z_0.

amateurs are connected to resonant antenna systems, which are principally resistive in nature. Consequently, the resistive load is an important practical case.

However, an antenna system is purely resistive only at one frequency, and when it is operated over a band of frequencies without readjustment—the usual condition—its impedance will contain a certain amount of reactance along with resistance. The effect of such a combination is to increase the standing-wave ratio—that is, as between two loads, one having only resistance of, say, 100 ohms as compared with a reactive load having the same total impedance, 100 ohms, the s.w.r. will be higher with the reactive load than with the purely resistive load. Also as between two loads containing the same value of resistance but one being without reactance while the other has a reactive component in addition, the s.w.r. will be higher with the one having the reactance.

The effect of reactance in the load is to shift the phase of the current with respect to the voltage both in the load itself and in the reflected components of voltage and current. This in turn causes a shift in the phase of the resultant current with respect to the resultant voltage. The net result is to shift the points along the line at which the various effects already described will occur. Representative cases for a load resistance less than the characteristic impedance of the line are shown in Fig. 3-17. With a load having inductive reactance the point of maximum voltage and minimum current is shifted toward the load. Comparing this with Fig. 3-12, it is as though the inductive reactance in the load of Fig. 3-17 had replaced a section (about 35 degrees in the illustration) of line, so that the line appears electrically longer than its actual length. The reverse occurs when the re-

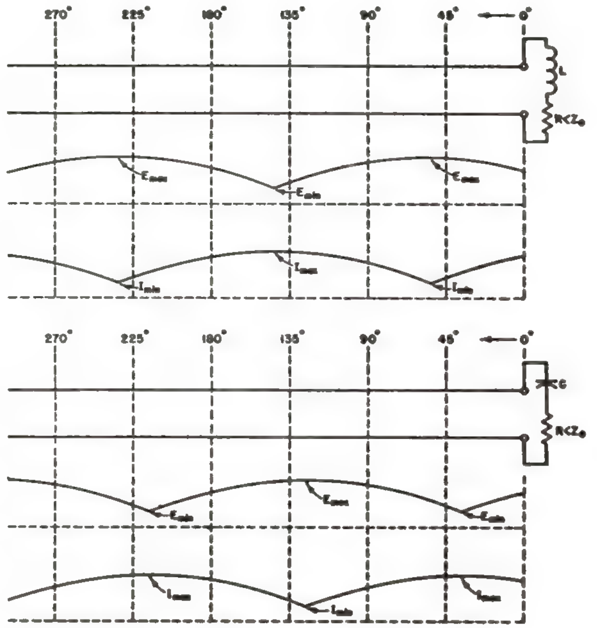


Fig. 3-17—Effect of reactance in the load. A reactive load causes the points of maximum voltage and minimum current to shift to a new position as compared with the case of a purely-resistive load. The shift is in opposite directions for inductive and capacitive loads. The cases shown are for *R* less than Z_0.

actance in the load is capacitive. This is shown in the lower drawings of Fig. 3-17; electrically, the line appears shorter than its actual length since more line would have to be added, if the capacitive reactance were not there, to bring the E_{max}-I_{min} point to its actual location.

When R is higher than Z_0 the E_{max}-I_{min} and E_{min}-I_{max} points are interchanged, just as in the case of purely resistive loads. The exact locations of these points depends on the ratio of reactance to resistance in the load. The higher this ratio the more the points are shifted as compared with their positions with purely-resistive loads.

ATTENUATION

The discussion in the preceding part of this chapter applies to all types of transmission lines, regardless of their physical construction. It is, however, based on the assumption that there is no power loss in the line. Every actual line will have some inherent loss, partly because of the resistance of the conductors, partly because of the fact that power is consumed in every dielectric used for insulating the conductors, and partly because in many cases a small amount of power escapes from the line by radiation.

Losses in a line modify its characteristic impedance slightly, but usually not to a sufficient extent to be significant. They will also affect the input impedance; in this case the theoretical values will be modified only slightly if the line is short and has only a small loss, but may be changed considerably if an appreciable proportion of the power input to the line is dissipated by the line itself. A large loss may exist because the line is long, because it has inherently high loss per unit length, because the standing-wave ratio is high, or because of a combination of two or all three of these factors.

The reflected power returning to the input terminals of the line is less when the line has losses than it would be if there were none. The over-all effect is that the s.w.r. changes along the line, being highest at the load and smallest at the input terminals. A long, high-loss line therefore tends to act, so far as its input impedance is concerned, as though the impedance match at the load end were better than is actually the case.

Line Losses

The power lost in a transmission line is not directly proportional to the line length but varies logarithmically with the length. That is, if 10% of the input power is lost in a section of line of certain length, 10% of the remaining power will be lost in the next section of the same length, and so on. For this reason it is customary to express line losses in terms of decibels per unit length, since the decibel is a logarithmic unit. Calculations are very simple because the total loss in a line is found by multiplying the db. loss per unit length by the total length of the line. Line loss is usually expressed in decibels per 100 feet. It is necessary to specify the frequency for which the loss applies, since the loss varies with frequency.

Conductor loss and dielectric loss both increase as the operating frequency is increased, but not in the same way. This, together with the fact that the relative amount of each type of loss depends on the actual construction of the line, makes it impossible to give a specific relationship between loss and frequency that will apply to all types of lines. Each line has to be considered individually. Actual loss values are given in a later section.

Effect of S.W.R.

The power lost in a line is least when the line is terminated in a resistance equal to its characteristic impedance, and increases with an increase in the standing-wave ratio. This is because the effective values of both current and voltage become larger as the s.w.r. becomes greater. The increase in effective current raises the ohmic losses in the conductors, and the increase in effective voltage increases the losses in the dielectric.

The increased loss caused by an s.w.r. greater than 1 may or may not be serious. If the s.w.r. at the load is not greater than 2, the *additional* loss caused by the standing waves, as compared with the loss when the line is perfectly matched, does not amount to more than about ½ db. even on very long lines. Since ½ db. is an undetectable change in signal strength, it can be said that from a practical standpoint an s.w.r. of 2 or less is, so far as losses are concerned, every bit as good as a perfect match.

The effect of s.w.r. on line loss is shown in Fig. 3-18. The horizontal axis is the attenuation, in decibels, of the line when perfectly matched. The vertical axis gives the *additional* attenuation, in decibels, caused by standing waves. For example, if the loss in a certain line is 4 db. when perfectly matched, an s.w.r. of 3 on that same line will cause an additional loss of 1.1 db., approximately. The total loss on the poorly-matched line is therefore $4 + 1.1 = 5.1$ db. If the s.w.r. were 10 instead of 3, the additional loss would be 4.3 db, and the total loss $4 + 4.3 = 8.3$ db.

It is of interest to note that when the line loss is high with perfect matching, the *additional* loss in db. caused by the s.w.r. tends to be constant regardless of the matched line loss. The reason for this is that the amount of power available to be reflected from the load is reduced, because relatively little power reaches the load in the first place. For example, if the line loss with perfect matching is 6 db., only 25% of the power originally put into the line reaches the load. If the mismatch at the load (the s.w.r. at the load) is 4 to 1, 36% of the power reaching the load will be reflected. Of the power originally put into the line, then, $0.25 \times 0.36 = 0.09$ or 9% will be reflected. This in turn will be attenuated 6 db. traveling back to the input end of the line, so that only $0.09 \times 0.25 = 0.0225$ or slightly over 2% of

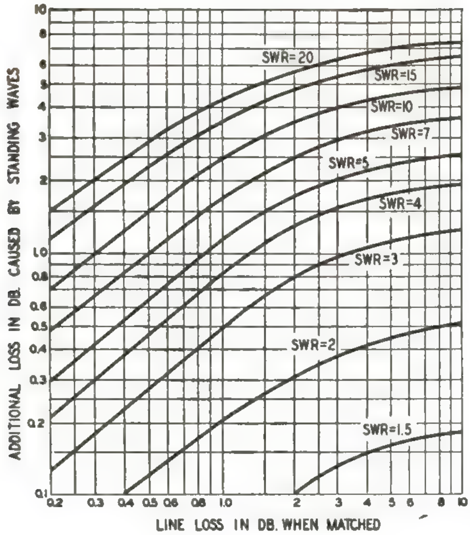


Fig. 3-18—Increase in line loss because of standing waves (s.w.r. measured at the load). To determine the total loss in decibels in a line having an s.w.r. greater than 1, first determine the loss for the particular type of line, length and frequency, on the assumption that the line is perfectly matched (Tables 3-1 and 3-11). Locate this point on the horizontal axis and move up to the curve corresponding to the actual s.w.r. The corresponding value on the vertical axis gives the additional loss in decibels caused by the standing waves.

the original power actually gets back to the input terminals. With such a small proportion of power returning to the input terminals the s.w.r. measured at the *input* end of the line would be only about 1.35 to 1—although it is 4 to 1 at the load. In the presence of line losses the s.w.r. always decreases along the line going from the load to the input end.

On lines having low losses when perfectly matched, a high standing-wave ratio may increase the power loss by a large factor. However, in this case the *total* loss may still be inconsequential in comparison with the power delivered to the load. An s.w.r. of 10 on a line having only 0.3 db. loss when perfectly matched will cause an additional loss of 1 db., as shown by the curves. This loss would produce a just-detectable difference in signal strength.

VOLTAGES AND CURRENTS ON LINES

The power reflected from a mismatched load does not represent an actual loss, except as it is attenuated in traveling back to the input end of the line. It merely represents power returned, and the actual effect is to reduce the power taken from the source. That is, it reduces the coupling between the power source and the line. This is easily overcome by readjusting the coupling until the actual power put into the line is the same as it would be with a matched load. In doing this, of course, the voltages and currents at loops along the line are increased.

As an example, suppose that a line having a characteristic impedance of 600 ohms is matched by a resistive load of 600 ohms and that 100 watts of power goes into the input terminals. The line simply looks like a 600-ohm resistance to the source of power. By Ohm's Law the current and voltage in such a matched line are

$$I = \sqrt{\frac{P}{R}}$$

$$E = \sqrt{PR}$$

Substituting 100 watts for P and 600 ohms for R, the current is 0.408 ampere and the voltage is 245 volts. Assuming for the moment that the line has no losses, all the power will reach the load so the voltage and current at the load will be the same as at the input terminals.

Now suppose that the load is 60 ohms instead of 600 ohms. The s.w.r. is 10, therefore. The reflection coefficient, or ratio of the reflected voltage or current to the voltage or current arriving at the load, is

$$k = \frac{S.W.R. - 1}{S.W.R. \times 1}$$

In this case the reflection coefficient is $(10 - 1)/(10 + 1) = 9/11 = 0.818$, so that the reflected voltage and current are both equal to 81,8% of the incident voltage and current. The reflected *power* is proportional to the square of either the current or voltage, and so is equal to $(0.818)^2 = 0.67$ times the incident power, or 67 watts. Since we have assumed that the line has no losses, this amount of power arrives back at the input terminals and subtracts from the original 100 watts, leaving only 33 watts as the amount of power actually taken from the source.

In order to put 100 watts into the 60-ohm load the coupling to the source must be increased so that the incident power minus the reflected power equals 100 watts, and since the power absorbed by the load is only 33% of that reaching it, the incident power must equal $100/0.33 = 303$ watts. In a perfectly-matched line, the current and voltage with 303 watts input would be 0.71 ampere and 426 volts, respectively. The reflected current and voltage are 0.818 times these values, or 0.581 ampere and 349 volts. At current maxima or loops the current will therefore be $0.71 + 0.58 = 1.29$ amp., and at a minimum point will be $0.71 - 0.58 = 0.13$ amp. The voltage maxima and minima will be $426 \times 349 = 775$ volts and $426 - 349 = 77$ volts. (Because of rounding off figures in the calculation process the s.w.r. does not work out to be exactly 10 in either the voltage or current case, but the error is very small.

In the interests of simplicity this example has been based on a line with no losses, but the approximate effect of line attenuation could be in-

cluded without much difficulty. If the matched-line loss were 3 db., for instance, only half the input power would reach the load, so new values of current and voltage at the load would be computed accordingly. The reflected power would then be based on the attenuated figure, and then itself attenuated 3 db. to find the power arriving back at the input terminals. The over-all result would be, as stated before, a reduction in the s.w.r. at the input terminals as compared with that at the load, along with less actual power delivered to the load for the same power input to the line.

Fig. 3-19 shows the ratio of current or voltage at a loop, in the presence of standing waves, to the current or voltage that would exist with the same power in a perfectly-matched line. Strictly speaking, the curve applies only near the load in the case of lines with appreciable losses. However, the curve shows the maximum possible value of current or voltage that can exist along the line whether there are line losses or not, and so is useful in determining whether or not a particular line can operate safely with a given s.w.r.

SPECIAL CASES

Besides the primary purpose of transporting power from one point to another, transmission lines have properties that are useful in a variety of ways. One such special case is a line an exact multiple of one-quarter wavelength (90 degrees) long. As shown earlier, such a line will have a purely-resistive input impedance when the termination is a pure resistance. Also, unterminated —i.e., short-circuited or open-circuited—lines can be used in place of conventional inductors and capacitors since such lines have an input impedance that is substantially a pure reactance when the line losses are low.

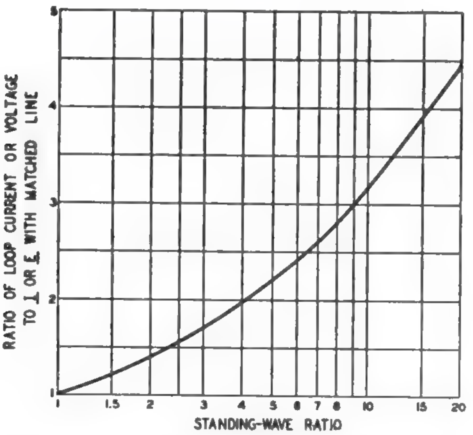


Fig. 3-19—Increase in maximum value of current or voltage on a line with standing waves, as referred to the current or voltage on a perfectly-matched line, for the same power delivered to the load. Voltage and current at minimum points are given by the reciprocals of the values along the vertical axis.

The Half-Wavelength Line

When the line length is an even multiple of 90 degrees (that is, a multiple of half wavelength), the input resistance is equal to the load resistance. As a matter of fact, a line an exact multiple of a half wave in length simply repeats, at its input or sending end, whatever impedance exists at its output or receiving end; it does not matter whether the impedance at the receiving end is resistive, reactive, or a combination of both. Sections of line having such length can be cut in or out without changing any of the operating conditions, at least when the losses in the line itself are negligible.

Impedance Transformation with Quarter-wave Lines

The input impedance of a line an odd multiple of a quarter wavelength long is

$$Z_s = \frac{Z_0^2}{Z_r}$$

Where Z_S is the input impedance and Z_R is the load impedance. If Z_R is a pure resistance, Z_S also will be a pure resistance. Rearranging this equation gives

$$Z_o = \sqrt{Z_s Z_r}$$

This means that if we have two values of impedance that we wish to "match," we can do so if we connect them together by a quarter-wave transmission line having a characteristic impedance equal to the square root of their product.

A quarter-wave line is, in effect, a transformer. It is frequently used as such in antenna work when it is desired, for example, to transform the impedance of an antenna to a new value that will match a given transmission line. This subject is considered in greater detail in a later section of this chapter.

Lines as Circuit Elements

An open- or short-circuited line does not deliver any power to a load, and for that reason is not, strictly speaking, a "transmission" line. However, the fact that a line of the proper length has inductive reactance makes it possible to substitute the line for a coil in an ordinary circuit. Likewise, another line of appropriate length having capacitive reactance can be substituted for a capacitor. The general way in which the reactance varies with line length is shown in Figs. 3-20 and 3-21. Note that the type of reactance exhibited at the input terminals of a line of given length depends on whether it is open- or short-circuited at the far end.

At lengths that are exact multiples of a quarter wavelength such lines have the properties of resonant circuits. At lengths where the input reactance passes through zero the line acts like a series-resonant circuit. At lengths for which the reactances theoretically pass from "positive" to

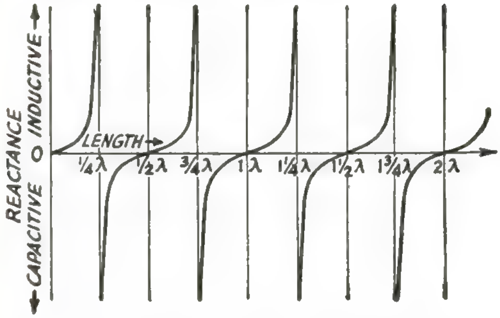


Fig. 3-20—Reactance at the input terminals as a function of line length in wavelengths, short-circuited line. These curves show the general way in which the input reactance varies. Specific values are determined by the characteristic impedance of the line.

"negative" infinity the line simulates a parallel-resonant circuit. The effective Q of such linear resonant circuits is very high if the line losses, both in resistance and by radiation, are kept down. This can be done without much difficulty, particularly in coaxial lines, if air insulation is used between the conductors. Air-insulated open-wire lines are likewise very good at frequencies for which the conductor spacing is very small in terms of wavelength.

Sections of lines used as circuit elements are usually a quarter wavelength or less long. The desired type of reactance (inductive or capacitive) or the desired type of resonance (series or parallel) is obtained by shorting or opening the far end of the line. The circuit equivalents of various types of line sections are shown in Fig. 3-22.

When a line section is used as a reactance, the amount of reactance obtained is determined by the characteristic impedance and the electrical length of the line. In the case of a line having no losses, and to a close approximation when the losses are small, the inductive reactance of a short-circuited line less than a quarter wave in length is

$$X_L \text{ (ohms)} = Z_0 \tan l$$

where l is the length of the line in electrical degrees and Z_0 is the characteristic impedance of the line. The capacitive reactance of an open-circuited line less than a quarter wave in length is

$$X_c \text{ (ohms)} = Z_0 \cot l$$

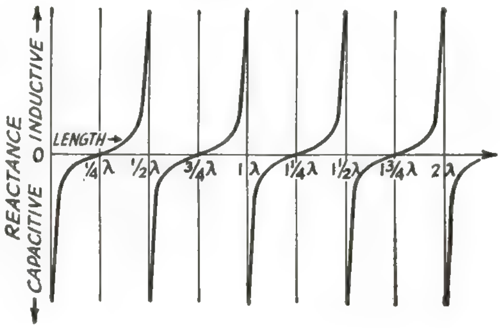


Fig. 3-21—Reactance at the input terminals as a function of line length in wavelengths open-circuited line.

Fig. 3-23 is a graph of the quantity X/Z_0 for both cases. To find the actual reactance at a given length, multiply the value of X/Z_0 by the characteristic impedance of the line used. For example, a section of 600-ohm line 60 degrees long and short-circuited at the far end will have an inductive reactance of $1.73 \times 600 = 1040$

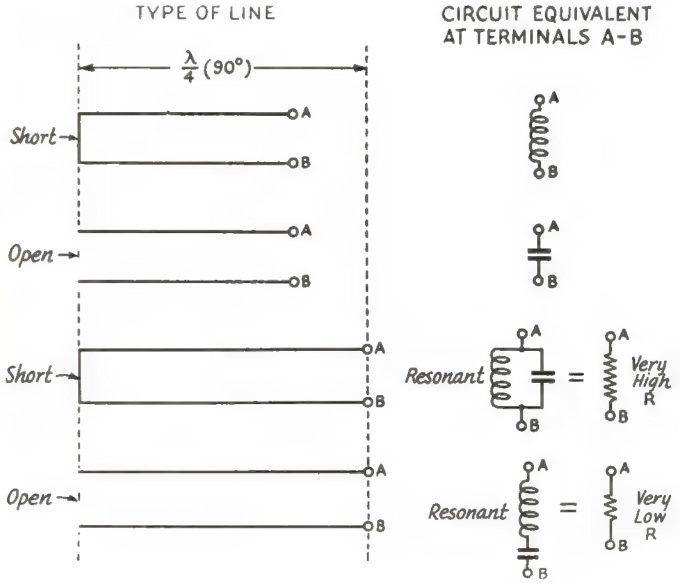


Fig. 3-22—Lumped-constant circuit equivalents of open- and short-circuited transmission lines.

ohms. The same line open-circuited would have a capacitative reactance of $0.57 \times 600 = 340$ ohms. The equivalent inductance and capacitance can be found by substituting these values in the formulas relating induitance and capacitance to reactance, or by using the various charts and calculators available. The frequency corresponding to the line length in degrees must be used, of course. In this example, if the frequency is 14 Mc. the equivalent inductance and capacitance in the two cases are 12.1 μh. and 33.4 $\mu\mu$f., respectively. Note that when the line length is 45 degrees (⅛ wavelength) the reactance in either case is numerically equal to the characteristic impedance of the line.

Applications of line sections as circuit elements in connection with antenna and transmission-line systems are discussed later in this chapter. In using the graphs in Fig. 3-23 it should be kept in mind that the electrical length depends on the frequency and velocity of propagation as well as on the actual physical length.

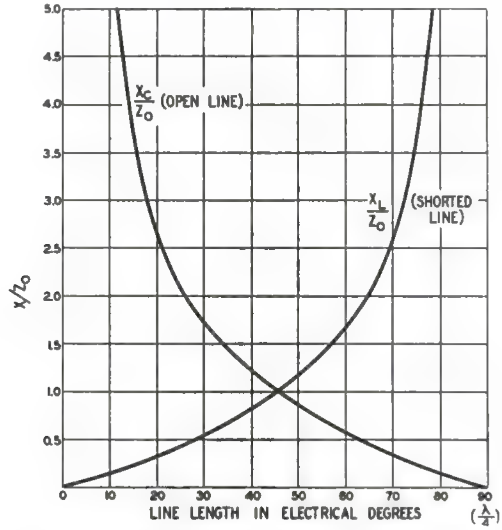


Fig. 3-23—Universal reactance curves for open- and short-circuited low-loss lines. The quantity X/Zo multiplied by the characteristic impedance of the line is equal to the value of reactance at the input terminals. In lines of greater length the type of reactance varies according to the following table:

Length	*Short-Circuited Line*	*Open-Circuited Line*
90 to 180 deg.	Capacitive	Inductive
180 to 270 deg.	Inductive	Capacitive
270 to 360 deg.	Capacitive	Inductive

The sign of the reactance reverses in each consecutive 90-degree (¼-wavelength) section.

Line Construction And Operating Characteristics

The two basic types of transmission lines, parallel-conductor and coaxial, can be constructed in a variety of forms. Both types can be divided into two classes: those in which the majority of the insulation between the conductors is air, only the minimum of solid dielectric necessary for mechanical support being used; and those in which the conductors are imbedded in and separated by a solid dielectric. The former class (air-insulated) has the lowest loss per unit length because there is no power loss in dry air so long as the voltage between conductors is below the value at which corona forms. At the maximum power permitted in amateur transmitters it is seldom necessary to consider corona unless the s.w.r. on the line is very high.

AIR-INSULATED LINES

A typical type of construction used for parallel-conductor or "two-wire" air-insulated transmission lines is shown in Fig. 3-24. The two line wires are supported a fixed distance apart by means of insulating rods called spacers. Spacers may be made from insulating material, such as bakelite, or can be purchased ready-made. Materials commonly used in manufactured spacers are isolantite, Lucite, and polystyrene. The spacings used vary from two to six inches, the smaller spacings being desirable at the higher frequencies (28 Mc.) so that radiation will be minimized. It is necessary to use the spacers at small-enough intervals along the line to prevent the two wires from swinging appreciably with respect to each other in a wind. For amateur purposes, lines using this construction ordinarily have No. 12 or No. 14 conductors, and the characteristic impedance is from 500 to 600 ohms. Although once in universal use, such lines have now been largely superseded by prefabricated lines.

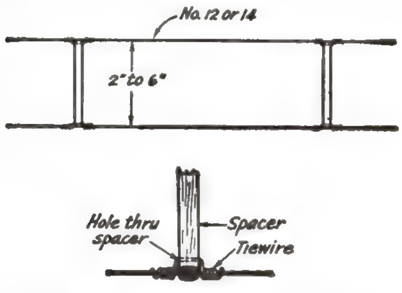


Fig. 3-24—Typical open-wire line construction. Commercial spacers are usually provided with grooved ends for the line conductors. The conductor is held in place by a tie wire anchored in a hole near the groove.

Prefabricated open-wire lines (sold principally for television receiving applications) are available in nominal characteristic impedances of 450 and 300 ohms. The spacers, of low-loss material such as polystyrene, are molded on the conductors at relatively small intervals so there is no tendency for the conductors to swing with respect to each other. A conductor spacing of one inch is used in the "450-ohm" line and ½ inch in the "300-ohm" line. The conductor size is usually about No. 18. The impedances of such lines are somewhat lower than given by Fig. 3-25 for the same conductor size and spacing, because of the effect of the dielectric constant of the numerous spacers used. The attenuation is quite low and lines of this type are entirely satisfactory for transmitting applications at amateur powers.

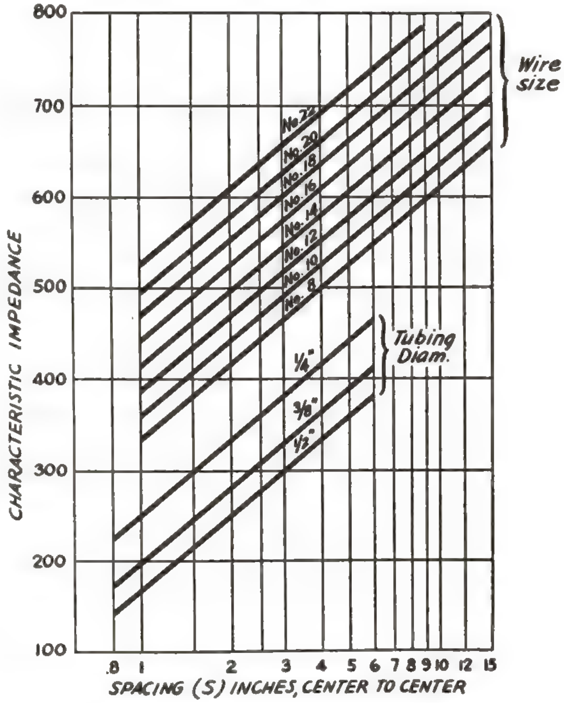


Fig. 3-25—Characteristic impedance vs. conductor size and spacing for parallel-conductor lines.

When an air-insulated line having still lower characteristic impedance is needed, metal tubing having a diameter from ¼ to ½ inch is frequently used. With the larger conductor diameter and relatively close spacing it is possible to build a line having a characteristic impedance as low as about 200 ohms. This type of construction is used principally for quarter-wave matching transformers at the higher frequencies.

Characteristic Impedance

The characteristic impedance of an air-insulated parallel-conductor line, neglecting the

effect of the insulating spacers, is given by:

$$Z_0 = 276 \log \frac{b}{a}$$

where Z_0 = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b

It does not matter what units are used for a and b so long as they are the *same* units. Both quantities may be measured in centimeters, inches, etc. Since it is necessary to have a table of common logarithms to solve practical problems, the solution is given in graphical form in Fig. 3-25 for a number of common conductor sizes.

Four-Wire Lines

Another type of parallel-conductor line that is useful in some special applications is the four-wire line. In cross-section, the conductors of the four-wire line are at the corners of a square, the spacings being of the same order as those used in two-wire lines. The conductor at opposite corners of the square are connected together to operate in parallel. This type of line has a lower characteristic impedance than the simple two-wire type. Also, because of the more symmetrical construction it is better balanced, electrically, to ground and other objects that may be close to the line. The spacers for a four-wire line may be disks of insulating material, "X"-shaped members, etc. The characteristic impedance of four-wire lines is shown graphically in Fig. 3-26.

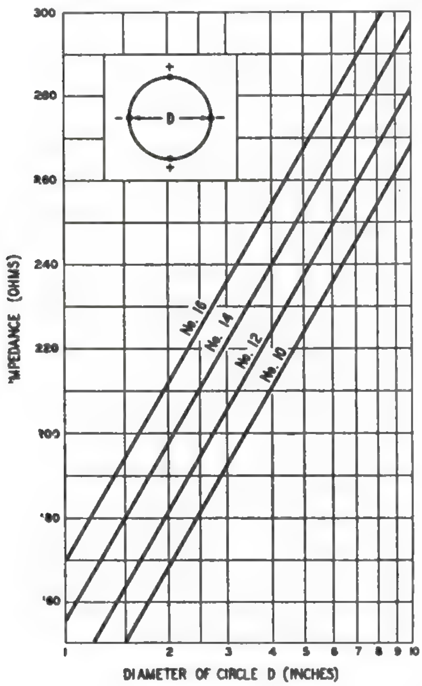


Fig. 3-26—Characteristic impedance vs. conductor size and spacing for four-wire lines. Opposite wires (not adjacent ones) are connected together at each end of the line.

Coaxial Lines

In coaxial lines of the air-insulated type a considerable proportion of the insulation between conductors may actually be a solid dielectric, because of the necessity for maintaining constant separators between the inner and outer conductors. This is particularly likely to be true in small-diameter lines, typical construction of which is shown in Fig. 3-27. The inner conductor, usually a solid copper wire, is supported by insulating beads at the center of the copper-tubing outer conductor. The beads usually are isolantite and the wire is generally crimped on each side of each bead to prevent the beads from sliding. The material of which the beads are made, and the number of them per unit length of line, will

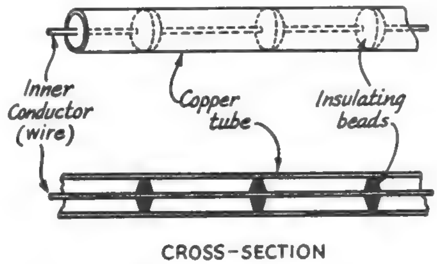


Fig. 3-27—Construction of air-insulated coaxial lines.

affect the characteristic impedance of the line. The greater the number of beads in a given length the lower the characteristic impedance compared with the value that would be obtained with air insulation only. The presence of the solid dielectric also increases the losses in the line. On the whole, however, a coaxial line of this type tends to have lower actual loss, at frequencies up to about 100 Mc., than any other line construction, provided the air inside the line can be kept dry. This usually means that air-tight seals must be used at the ends of the line and at every joint.

The characteristic impedance of an air-insulated coaxial line is given by the formula

$$Z_0 = 138 \log \frac{b}{a}$$

where Z_0 = Characteristic impedance
b = Inside diameter of outer conductors
a = Ontside diameter of inner conductor (in same units as b)

Again it does not matter what units are used for b and a, so long as they are the same. Curves for typical conductor sizes are given in Fig. 3-28.

The formula and curves for coaxial lines are approximately correct for lines in which bead

spacers are used, provided the beads are not too closely spaced.

FLEXIBLE LINES

Coaxial lines in which the conductors are separated by a flexible dielectric have a number of advantages over the air-insulated type. They are

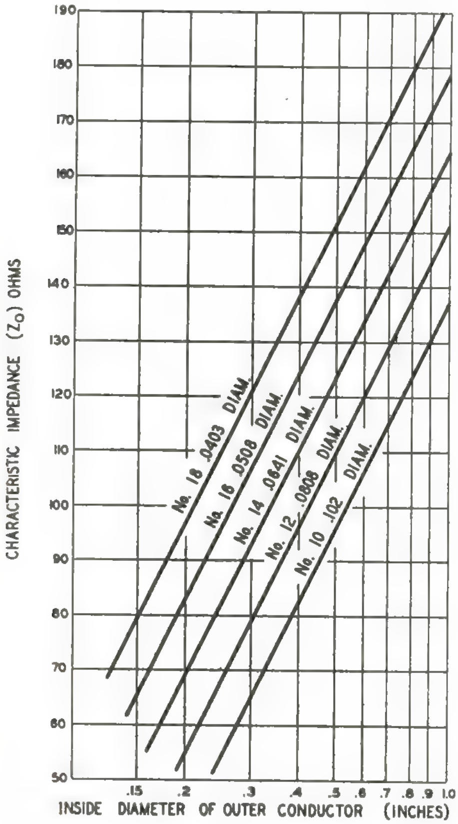


Fig. 3-28—Characteristic impedance of typical air-insulated coaxial lines.

less bulky, weigh less in comparable types, maintain more uniform spacing between conductors, are generally easier to install, and are neater in appearance. Both parallel-conductor and coaxial lines are available with this type of insultion.

The chief disadvantage of such lines is that the power loss per unit length is greater than in air-insulated lines. The power loss causes heating of the dielectric, and if the heating is great enough—as it may be with high power and a high standing-wave ratio—the line may break down both mechanically and electrically.

Parallel-Conductor Lines

The construction of a number of types of flexible lines is shown in Fig. 3-29. In the most common 300-ohm type ("Twin-Lead" is one trade name) the conductors are stranded wire equivalent to No. 20 in cross-sectional area and are molded in the edges of a polyethylene ribbon about a half inch wide. The effective dielectric is partly solid and partly air. The presence of the solid dielectric lowers the characteristic impedance of the line as compared with the same conductors in air, the result being that the impedance is approximately 300 ohms. The fact that part of the field between the conductors exists outside the solid dielectric leads to an operating disadvantage in that dirt or moisture on the surface of the ribbon tends to change the characteristic impedance. The operation of the line is therefore affected by weather conditions. The effect will not be very serious in a line terminated in its characteristic impedance, but if there is a considerable standing-wave ratio a small change in Z_0 may cause wide fluctuations of the input impedance. Weather effects can be minimized by cleaning the line occasionally and giving it a thin coating of a water-repellent material such as silicone grease or automobile wax.

To overcome the effects of weather on the characteristic impedance and attenuation of ribbon-type line, another type of Twin-Lead is made using a polyethylene tube with the conductors molded diametrically opposite each other in the walls. This increases the leakage path across the dielectric surface. Also, much of the electric field between the conductors is in the hollow center of the tube, which can be protected from dirt and weather by closing the exposed end of the tube to make it watertight. This type of line is fairly impervious to weather effects. Care should be used when installing it, however, to make sure that any moisture that condenses on the inside with changes in temperature and humidity can drain out the bottom end of the tube and not be trapped in one section. This type of line is made in two conductor sizes (with different tube diameters), one for receiving applications and the other for transmitting.

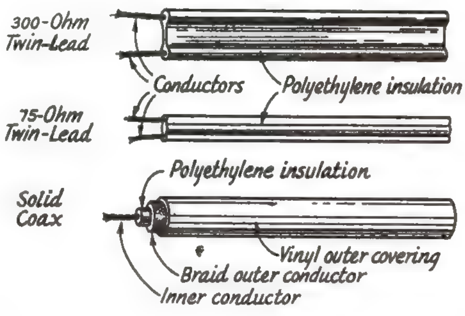


Fig. 3-29—Construction of parallel-conductor and coaxial lines with solid dielectric.

The transmitting-type 75-ohm Twin-Lead uses stranded conductors about equivalent to solid No. 12 wire, with quite close spacing between conductors. Because of the close spacing most of the field is confined to the solid dielectric, very little existing in the surrounding air. This makes the 75-ohm line much less susceptible to weather effects than the 300-ohm ribbon type.

TABLE 3-I—TRANSMISSION LINES

	1	2	3	4	5	6	7	8	9	10	11		12
Type of Line	*Nominal Imp., ohms*	*RG/U Type*	*Mfrs. No.*	*Outside Diam., inches*	*Jacket*	*Inner Cond. Size*	*Dielectric*	*Cap. per ft., pf.*	*Velocity Factor*	*Max. r.m.s. Voltage*	*Power Rating, Watts Up to*		*Connector Series*
											30 MHz.	*400 MHz.*	
Flexible Coaxial Medium	52	8		.405	I	7/21	SP	29.5	.66	5000	1720	465	UHF, N
	52	8A		.405	IIA	7/21	SP	29.5	.66	5000	1720	465	UHF, N
	50		621-111	.405	I	7/19	FP	24.5	.80	1500			UHF
	50		T-4-50	.407	X	10	FP						UHF
	75	11		.405	I	7/26	SP	20.5	.66	5000	1400	340	UHF, N
	75	11A		.405	IIA	7/26	SP	20.5	.66	5000	1400	340	UHF, N
	75		621-100	.405	I	14	FP	16.5	.80	3000			UHF
	75		JT-204	.407	X	14	FP						UHF
Small	53.5	58		.195	I	20	SP	28.5	.66	1900	580	135	UHF, BNC, N
	50	58A		.195	I	19/.0071	SP	30	.66	1900	550	105	UHF, BNC, N
	53.5	58B		.195	IIIA	20	SP	28.5	.66	1900	580	135	UHF, BNC, N
	50	58C		.195	IIA	19/.0071	SP	30	.66	1900	550	105	UHF, BNC, N
	73	59		.242	I	22 cw	SP	21.5	.66	2300	720	185	UHF, BNC, N
	75	59B		.242	IIA	.023 cw	SP	21	.66	2300	720	185	UHF, BNC, N
	73		621-186	.242	P	20 cw	FP	17.3	.80	1000			UHF, BNC
	93	62		.242	I	22 cw	SSP	13.5	.84	750	850	230	UHF, BNC, N
Parallel Conductor Flat or Oval	75		214-023			7/21	SP	20	.71		1000		
	300		214-056			7/28	SP	5.8	.82				
	300		214-022			16 cw	SP	3.0	.82				
Tubular	300		214-271			7/28	PA		.82		500		
	300		214-076			7/26	PA	3.9	.82		1000		
	300		214-103			7/28	FP*						

Column 3: T-4-50 and JT-204 are manufactured by Times Wire & Cable, Wallingford, Conn. Other numbers are types made by Amphenol, Chicago, Ill.

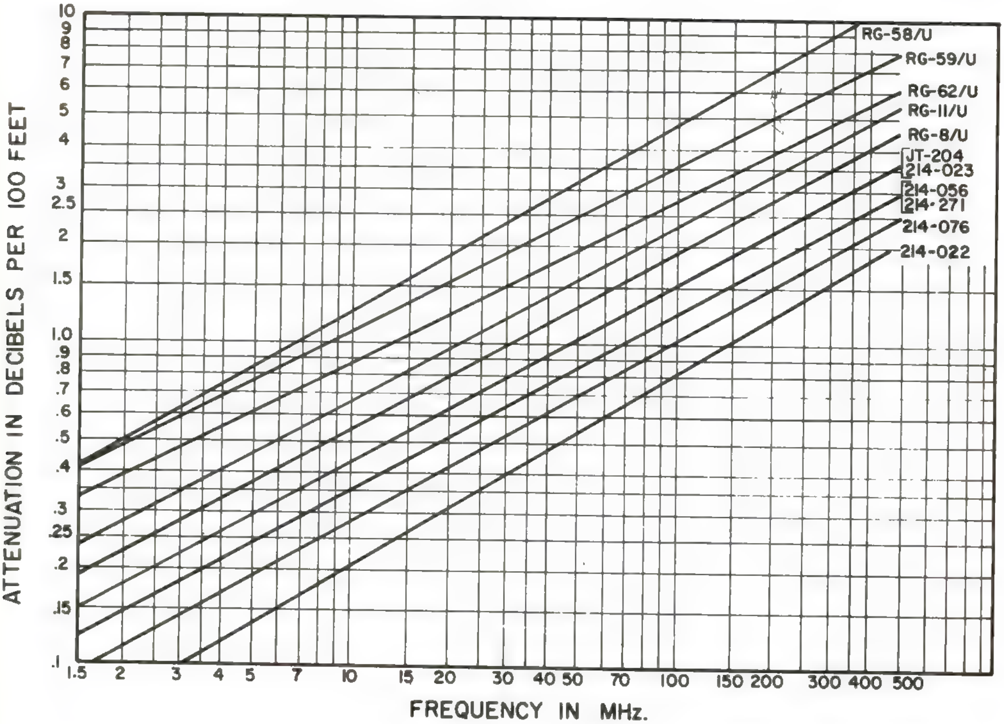
Column 5: I—Polyvinyl chloride (PVC), black. IIA—Noncontaminating PVC, black or gray. IIIA—Polyethylene, black. Noncontaminating and abrasion-resistant. Recommended when cable is to be buried underground. P—Polyethylene. X—Xelon.

Column 6: Conductors are copper unless followed by CW (copper-weld). Decimal numbers give wire diameter in inches; others are standard copper-wire gauge except when preceding a slant bar, when the figure indicates number of strands: e.g., 7/21 means 7 strands of No. 21 copper wire.

Column 7: SP—Solid polyethylene. SSP—Polyethylene strand wound around inner conductor; enclosed in solid tube of same material. FP—Foamed polyethylene. FP*—Foamed polyethylene surrounding each conductor; outer enclosure solid polyethylene. Type 214-103 is intended for use under adverse moisture and salt-spray conditions. PA—Polyethylene tube with air core.

Column 12: Only connectors in common use by amateurs are included.

Open parallel-conductor line has a velocity factor of 0.95 to 0.975, depending on number of spacers and dielectric material of which they are made.

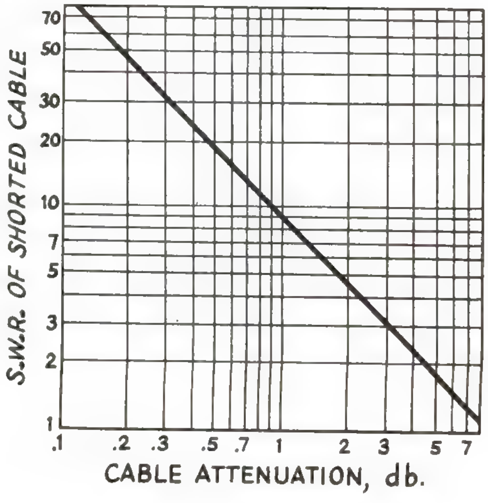


Nominal attenuation in decibels per 100 feet of various types of transmission line. Total attenuation is directly proportional to length. Attenuation will vary somewhat in actual cable samples, and generally increases with age in coaxial cables having a type I jacket. Cables grouped together in the above chart have approximately the same attenuation.

Types having foamed polyethylene dielectric have slightly lower loss than equivalent solid-dielectric types, when not specifically shown above. The curve for RG-58/U also applies to RG-58B/U. For RG-58A/U and RG-58C/U add 10 percent to the loss in decibels given by the curve. RG-8A/U has the same loss as RG-8/U; RG-59B/U has the same loss as RG-59.

TESTING OLD COAXIAL CABLE

Unknown coaxial cable or cable that has been exposed to the weather may have losses above the published figures for the cable type. If one has access to a sensitive s.w.r. bridge, the cable can be checked for losses at the frequency to be used. Connect the cable to the bridge and a low-powered source of r.f., and short circuit the far end of the cable. The s.w.r. measurement can then be transformed to the line loss (when perfectly terminated) by referring to the chart at the right.



By short-circuiting the far end of a length of transmission line and measuring the s.w.r. at the transmitter end, the loss in the line (when perfectly terminated) can be found from this chart.

In addition to the parallel-conductor types described above, there are also lightweight two-wire lines of 150 and 75 ohms. These are useful for receiving antennas, but are not heavy enough to carry very much power.

Coaxial Lines

Flexible "coax" is available in a rather large number of different types. However, the basic construction is the same in all, and is typified in the drawing in Fig. 3-29. The over-all diameter varies from a little less than ¼ inch to somewhat over an inch, depending chiefly on the power requirements for which the cable was designed. In some cables the inner conductor is stranded; in others, solid wire is used. In some the outer conductor is a single braid; in others it is double. The outer jacket, usually vinyl plastic plays no part in the electrical performance of the cable but is simply a waterproof covering.

The dielectric material used in these cables is polyethylene, a flexible plastic having low losses at radio frequencies. In many types of cable (particularly the RG series) the dielectric is solid polyethylene. Although the losses in the solid dielectric are relatively low, if the operating frequency is not too high, methods have been used to reduce them still further. A few types of cable (RG-62/U is an example) use a polyethylene thread wound around the inner conductor; this makes the insulation partly air, effecting a reduction in both loss and capacitance per unit length. A popular type of dielectric in the more-recently-developed cables is "foamed" polyethylene, which has a cellular structure similar to a honeycomb so that a large part of the dielectric is air. The loss per unit length in such cables is appreciably less than in those using solid dielectric.

Solid coaxial cables are available in two principal characteristic impedances, approximately 50 and 75 ohms.

If the entire field between the two conductors of a line is in a solid dielectric, as in the case of solid coaxial lines, the characteristic impedance of the line is reduced by the factor $1/\sqrt{k}$ as compared with the impedance of an air-insulated line having the same conductor size and spacing. The quantity k is the effective dielectric constant of the insulating material. In ribbon or tubular type parallel-conductor lines and in the special coaxial types mentioned above, the field is partly in air and partly in the dielectric, so the reduction factor above cannot be applied directly.

The attenuation and other characteristics of the various types of lines commonly used by amateurs are shown in Table 3-I.

A whole series of fittings for making detachable connections to flexible coaxial cable is available These include general-purpose connectors, some of which are quite inexpensive, and "constant-impedance" units especially designed so that lengths of cable can be spliced together or terminated without causing a change in the characteristic impedance. Such impedance "bumps" along a line correspond in a general way to having a load that is not matched to the line; that is, they will cause some of the outgoing power to be reflected back toward the input end. In most amateur applications it is not necessary to worry about such impedance discontinuities when using ordinary connectors because their effect at frequencies below 300 or 400 Mc. is too small to be of practical consequence.

SINGLE-WIRE LINE

There is one type of line, in addition to those already described, that deserves some mention since it is still used to a limited extent. This is the **single-wire line**, consisting simply of a single conductor running from the transmitter to the antenna. The "return" circuit for such a line is the earth; in fact, the second conductor of the line can be considered to be the image of the actual conductor in the same way that an antenna strung above the earth has an image (see Chapter Two). The characteristic impedance of the single-wire line depends on the conductor size and the height of the wire above ground, ranging from 500 to 500 ohms for No. 12 or No. 14 conductors at heights of 10 to 30 feet. By connecting the line to the antenna at a point that represents a resistive impedance of 500 to 600 ohms the line can be matched and will operate without standing waves.

Although the single-wire line is very simple to install, it has at least two outstanding disadvantages. Since the return circuit is through the earth, the behavior of the system depends on the kind of ground over which the antenna and transmission line are erected. In practice, it may not be possible to get the necessary good connection to actual ground that is required at the transmitter. Second, the line always radiates since there is no near-by second conductor to cancel the fields. The radiation will be minimum when the line is properly terminated because the line current is lowest under those conditions. However, the line is always a part of the radiating antenna system to a greater or lesser extent.

ELECTRICAL LENGTH

Whenever reference is made to a line as being so many wavelengths (such as a "half wavelength" or "quarter wavelength") long, it is to be understood that the *electrical* length of the line is meant. The physical length corresponding to an electrical wavelength is given by

$$\textbf{Length (feet)} = \frac{984}{f} V$$

where f = Frequency in megacycles
V = Velocity factor

The **velocity factor** is the ratio of the actual velocity along the line to the velocity in free space. Values of V for several common types of lines are given in Table 3-I.

Because a quarter-wavelength line is frequently used as an impedance transformer, it is

convenient to calculate the length of a quarter-wave line directly. The formula is

$$\textbf{Length (feet)} = \frac{246}{f} V.$$

LINE INSTALLATION

One great advantage of coaxial line, particularly the flexible dielectric type, is that it can be installed with almost no regard for its surroundings. It requires no insulation, can be run on or in the ground or in piping, can be bent around corners with a reasonable radius, and can be "snaked" through places such as the space between walls where it would be impracticable to use other types of lines. However, coax lines always should be operated in systems that permit a low standing-wave ratio, and precautions must be taken to prevent r.f. currents from flowing on the *outside* of the line. This point is discussed later in this chapter.

Coaxial Fittings

There is a wide variety of fittings and connectors designed to go with various sizes and types of solid-dielectric coaxial line. The "u.h.f." series of fittings is by far the most widely-used type in the amateur field, largely because they have been available for a long time and have been quite inexpensive on the surplus market. These fittings, typified by the PL-259 plug and SO-239 chassis fitting (Armed Services numbers) are quite adequate for v.h.f. and lower-frequency applications, but are not weatherproof.

The "N" series fittings are designed to maintain constant impedance at cable joints. They are a bit harder to assemble than the "u.h.f." type, but are better for frequencies above 300 Mc. or so. These fittings are weatherproof.

The "BNC" fittings are for small cable such as RG-58/U, RG-59/U and RG-62/U. They feature a bayonet-locking arrangement for quick connect and disconnect, and are weatherproof.

Methods of assembling the connectors to the cable are shown on accompanying pages.

Parallel-Wire Lines

In installing a parallel-wire line care must be used to prevent it from being affected by moisture, snow and ice. In home construction, spacers that are impervious to moisture and are unaffected by sunlight and the weather should be used on air-insulated lines. Steatite spacers meet this requirement adequately, although they are somewhat heavy. The wider the line spacing the longer the leakage path across the spacers, but this cannot be carried too far without running into line radiation, particularly at the higher frequencies. Where an open-wire line must be anchored to a building or other structure, stand-off insulators of a height comparable with the line spacing should be used if mounted in a spot that is open to the weather. Lead-in bushings for bringing the line into a building also should have a long leakage path.

The line should be kept away from other conductors, including downspouting, metal window frames, flashing, etc., by a distance equal to two or three times the line spacing. Conductors that are very close to the line will be coupled to it in greater or lesser degree, and the effect is that of placing an additional load across the line at the point where the coupling occurs. Reflections take place from this coupled "load," raising the standing-wave ratio. The effect is at its worst when one line wire is closer than the other to the external conductor. In such a case one wire carries a heavier load than the other, with the result that the line currents are no longer equal. The line then becomes "unbalanced."

Solid-dielectric two-wire lines have a relatively small external field because of the small spacing and can be mounted within a few inches of other conductors without much danger of coupling between the line and such conductors. Stand-off insulators are available for supporting lines of this type when run along walls or similar structures.

Sharp bends should be avoided in any type of transmission line, because such bends cause a change in the characteristic impedance. The result is that reflections take place from each bend. This is of less importance when the s.w.r. is high than when an attempt is being made to match the load to the line's characteristic impedance. It may be impossible to get the s.w.r. down to a desired figure until the necessary bends in the line are made more gradual.

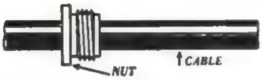
COUPLING THE TRANSMITTER TO THE LINE

In any system using a transmission line to feed the antenna the load that the transmitter "sees" is the input impedance of the line. As shown earlier, this impedance is completely determined by the line length, the Z_0 of the line, and the impedance of the load (the antenna) at the output end of the line. The line length and Z_0 are generally matters of choice regardless of the type of antenna used. The antenna impedance, which may or may not be known accurately, is (with Z_0) the factor that determines the standing-wave ratio.

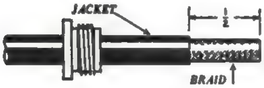
The s.w.r. can be measured with relative ease, and from it the limits of variation in the line input impedance can be determined with little difficulty. It may be said, therefore, that the problem of transferring power from the transmitter to the line can be approached purely on the basis of the known Z_0 of the line and the maximum s.w.r. that may be encountered.

So far as the transmitter itself is concerned, the requirements of present-day amateur operation almost invariably include complete shielding and provision for the use of low-pass filters

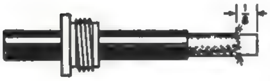
BNC Connectors



1.—Cut end of cable even.

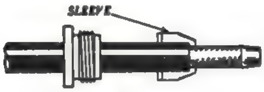


2.—Remove vinyl jacket ½"—don't nick braid.

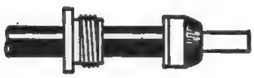


3.—Push braid back and remove ⅛" of insulation and conductor.

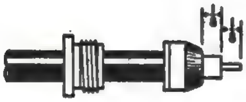
4.—Taper braid.



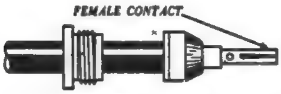
5.—Slide sleeve over tapered braid. Fit inner shoulder of sleeve squarely against end of jacket.



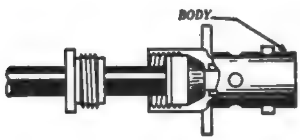
6. — With sleeve in place, cobb out braid, fold back smooth as shown, and trim 3/32".



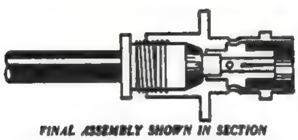
7.—Bare center conductor ⅛"—don't nick conductor.



8.—Tin center conductor of cable. Slip female contact in place and solder. Remove excess solder. Be sure cable dielectric is not heated excessively and swollen so as to prevent dielectric entering body.

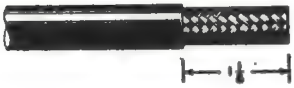


9.—Push into body as far as it will go. Slide nut into body and screw into pace, with wrench, until moderately tight. Hold cable and shell rigidly and rotate nut.

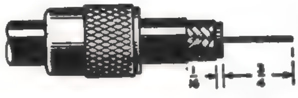


10.—This assembly procedure applies to BNC jacks. The assembly for plugs is the same except for the use of male contacts and a plug body.

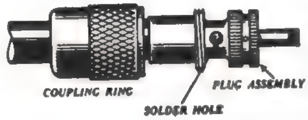
83-1SP (PL-259) Plug



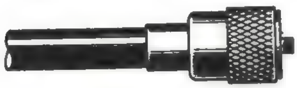
1.—Cut end of cable even. Remove vinyl jacket 1⅛"—don't nick braid.



2.—Bare ¾" of center conductor—don't nick conductor. Trim braided shield 9/16" and tin. Slide coupling ring on cable.



3.—Screw the plug assembly on cable. Solder plug assembly to braid through solder holes. Solder conductor to contact sleeve.

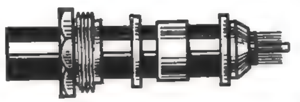


4.—Screw coupling ring on assembly.

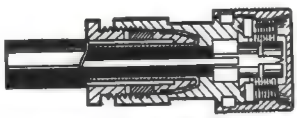
N Connectors

1.—Remove ½" of vinyl jacket. When using double shielded cable remove 9/16".

2.—Comb out copper braid as shown. Cut off dielectric ¼" from end. Tin center conductor.



3.—Taper braid as shown. Slide nut, washer and gasket over vinyl jacket. Slide clamp over braid with internal shoulder of clamp flush against end of vinyl jacket. When assembling connectors with gland, be sure knife-edge is toward end of cable and groove in gasket is toward the gland.



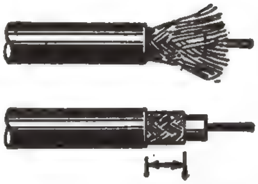
4.—Slide body into place carefully so that contact enters hole in insulator. Face of dielectric must be flush against insulator. Slide completed assembly into body by pushing nut. When nut is in place, tighten with wrenches. In connectors with gland, knife-edge should cut gasket in half by tightening sufficiently.

This material courtesy Amphenol Connector Division, Amphenol-Borg Electronics Corp.

83 Series (SO-239) with Hoods

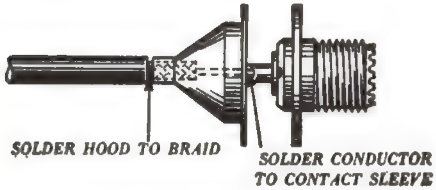
1.—Cut end of cable even. Remove vinyl jacket. Do not nick braid.

Remove braid and dielectric to expose center conductor. Do not nick conductor.

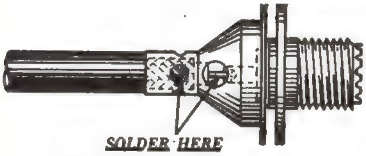


3.—Fan braided shielding.

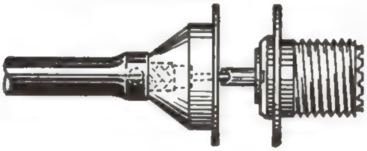
4.—Smooth out shielding Trim braid to size. Tin braid and wipe clean. Soldering and assembly depends on the hood used, as illustrated.



5.—Slide hood over braid. Solder conductor to contact. Slide hood flush against receptacle and bolt both to chassis. Solder hood to braid as illustrated. Tape this junction if necessary.

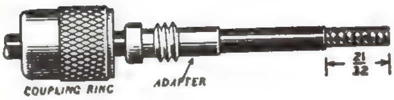


6.—Slide hood over braid. Bring receptacle flush against hood. Solder hood to braid and conductor to contact sleeve through solder holes as illustrated. Tape junction if necessary.



7.—Slide hood over braid and force under vinyl. Place inner conductor in contact sleeve and solder. Push hood flush against receptacle. Solder hood to braid through solder holes. Tap junction if necessary.

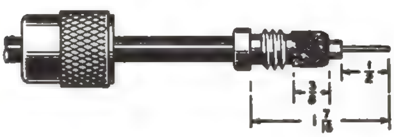
83-1SP (PL-259) Plug with Adapters



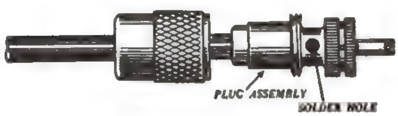
1.—Cut end of cable even. Remove vinyl jacket 21/32"—don't nick braid. Slide coupling ring and adapter on cable.

2.—Fan braid slightly and fold back over cable.

3.—Compress braid around cable. Position adapter to dimension shown. Press braid down over body of adapter to dimension shown. Press braid down over body of adapter and trim.



4.—Bare ½" of center conductor—don't nick conductor. Pre-tin exposed center conductor.



5.—Screw the plug assembly on adapter. Solder braid to shell through solder holes.
Solder conductor to contact sleeve.

6.—Screw coupling ring on back shell.

to prevent harmonic interference with television reception. In almost all cases this means that the use of coaxial cable at the output of the transmitter is mandatory because it is inherently shielded. Thus the transmitter output circuit should be designed to deliver full power into 50-75 ohms, with enough leeway to take care of reasonable variations in load—equivalent to the line input-impedance range associated with a 2-to-1 or perhaps as high as 3-to-1 s.w.r. in the line. This does not mean that a coaxial line must be used to feed the antenna, any type of line can be used.

The Matching System

The basic system then is as shown in Fig. 3-30. Assuming that the transmitter is capable of delivering its rated power into a load of the order of 50 to 75 ohms, the coupling problem is one of designing a matching circuit that will transform the actual line input impedance into a resistance of 50 or 75 ohms. This resistance will be unbalanced—that is, one side will be grounded—but the actual line to the antenna probably will be balanced (parallel-conductor line) when a matching circuit is required. Unbalanced (co-axial) line should not require a matching circuit, since it should be used only when it is sufficiently well matched by the antenna to operate at a low s.w.r.—one well within the coupling capabilities of the transmitter itself.

Several types of matching circuits are available. With some it is necessary to know the actual line input impedance with fair accuracy in order to arrive at a proper design. This information is not essential with the inductively-coupled circuit described in the next section, since it is capable of adjustment over a wide range.

depends on the effective Q of the circuit $L_1C_1R_1$, the reactance of L_2 at the operating frequency, and the coefficient of coupling, k, between the two coils. The approximate relationship is (assuming C_1 is properly tuned)

$$R_2 = k^2 X_{L2} Q.$$

where X_{L2} is the reactance of L_2 at the operating frequency. The value of L_2 is optimum when $X_{L2} = R_2$, in which case the desired value of R_2 is obtained when

$$k = \frac{1}{\sqrt{Q}}$$

This means that the desired value of R_2 may be obtained by adjusting either the coupling, k, between the two coils, or by changing the Q of the circuit $L_1C_1R_1$—or, if necessary, by doing both. If the coupling is fixed, as is often the case, Q must be adjusted to attain a match. Note that increasing the value of Q is equivalent to tightening the coupling, and vice versa.

If L_2 does not have the optimum value, the match may still be achieved by adjusting k and Q, but one or the other—or both—must have a larger value than is needed when X_{L2} is equal to R_2. In general, it is desirable to use as low a value of Q as is practicable, since low Q values mean that the circuit requires little or no readjustment when shifting frequency within a band (provided R_1 does not vary appreciably with frequency).

Circuit Arrangements

In Fig. 3-31A, Q is equal to R_1 in ohms divided by the reactance of C_1 in ohms, assuming L_1C_1 to be tuned to the operating frequency. This circuit is suitable for comparatively high

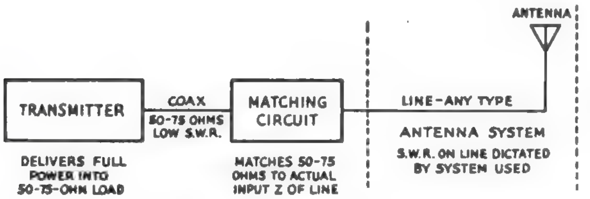


Fig. 3-30—Essentials of coupling system between transmitter and transmission line.

MATCHING WITH INDUCTIVE COUPLING

Inductively-coupled matching circuits are shown in basic form in Fig. 3-31. R_1, is the actual load resistance to which the power is to be delivered, and R_2 is the resistance seen by the power source. R_2 depends on the circuit design and adjustment; in general, the object is to make it equal to 50 or 75 ohms. L_1 and C_1 form a resonant circuit capable of being tuned to the operating frequency. The coupling between L_1 and L_2 is adjustable.

The circuit formed by C_1, L_1 and L_2 is equivalent to a transformer having a primary-to-secondary impedance ratio adjustable over wide limits. The resistance "coupled into" L_2 from L_1

values of R_1—from several hundred to several thousand ohms. In Fig. 3-31C, Q is equal to the reactance of C_1 divided by the resistance of R_1, L_1C_1 again being tuned to the operating frequency. This circuit is suitable for low values of R_1—from a few ohms up to a hundred or so ohms. In Fig. 3-31B the Q depends on the placement of the taps on L_1 as well as on the reactance of C_1. This circuit is suitable for matching all values of R_1 likely to be encountered in practice.

Note that to change Q in either A or C, Fig. 3-31, it is necessary to change the reactance of C_1. Since the circuit is tuned essentially to resonance at the operating frequency, this means that the L/C ratio must be varied in order to change Q.

In Fig. 3-31B a fixed L/C ratio may be used since Q can be varied by changing the tap positions. The Q will increase as the taps are moved closer together, and will decrease as they are moved farther apart on L_1.

Reactive Loads

More often than not, the load represented by the input impedance of the transmission line is reactive as well as resistive. In such a case the load cannot be represented by a simple resistance such as R_1 in Fig. 3-31. As stated earlier in this chapter, we have the option of considering the load to be a resistance in parallel with a reactance, or as a resistance in series with a reactance

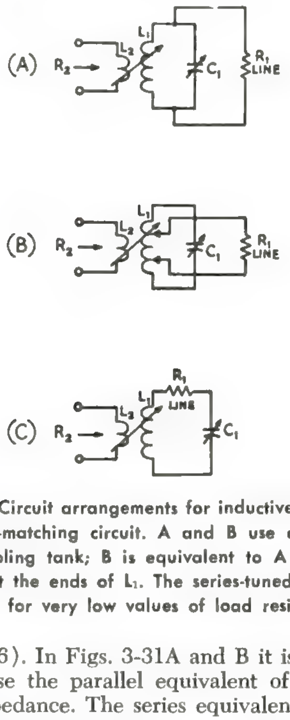


Fig. 3-31—Circuit arrangements for inductively-coupled impedance-matching circuit. A and B use a parallel-tuned coupling tank; B is equivalent to A when the taps are at the ends of L_1. The series-tuned circuit at C is useful for very low values of load resistance, R_1.

(Fig. 3-16). In Figs. 3-31A and B it is convenient to use the parallel equivalent of the line input impedance. The series equivalent is more suitable for Fig. 3-31C.

Thus in Fig. 3-32 at A and B the load might be represented by R_1 in parallel with the capacitive reactance C, and in Fig. 3-32C by R_1 in series with a capacitive reactance C. In A, the capacitance C is in parallel with C_1 and so the total capacitance is the sum of the two. This is the effective capacitance that, with L_1, tunes to the operating frequency. Obviously the setting of C_1 will be at a lower value of capacitance with such a load than it would with a purely resistive load such as is shown in Fig. 3-31A.

In Fig. 3-32B the capacitance of C also increases the total capacitance effective in tuning the circuit. However, in this case the increase in effective tuning capacitance depends on the positions of the taps; if the taps are close together the effect of C on the tuning is relatively small, but it increases as the taps are moved farther apart.

In Fig. 3-32C, the capacitance C is in series with C_1 and so the total capacitance is less than either. Hence the capacitance of C_1 has to be increased in order to resonate the circuit, as compared with the purely resistive load shown in Fig. 3-31C.

If the reactive component of the load impedance is inductive, similar considerations apply. In such case an inductance would be substituted for the capacitance C shown in Fig. 3-32. The effect in Figs. 3-32A and B would be to decrease the effective inductance in the circuit, so C_1 would have to have a larger value of capacitance in order to resonate the circuit to the operating frequency. In Fig. 3-32C the effective inductance would be increased, thus making it necessary to set C_1 at a lower value of capacitance for resonating the circuit.

Effect of Line Reactance on Circuit Q

The presence of reactance in the load can affect the Q of the matching circuit. If the reactance is capacitive, the Q will not change if resonance can be maintained, by adjustment of C_1, without changing the value of L_1 (or the position of the taps in Fig. 3-32B), as compared with the Q when the load is purely resistive and has the same value of resistance, R_1. If the load reactance is inductive the L/C ratio changes because the effective inductance in the circuit is changed and, in the ordinary case, L_1 is not adjustable. This increases the Q in all three circuits.

When the load has appreciable reactance it is not always possible to adjust the circuit to resonance by readjusting C_1, as compared with the setting it would have with a purely resistive load. Such a situation may occur when the load

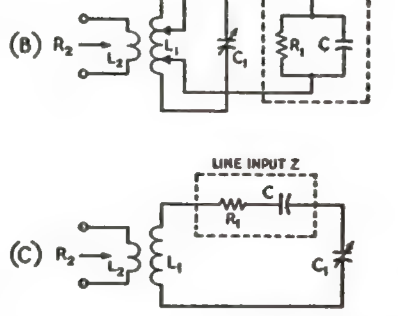


Fig. 3-32—Line input impedances containing both resistance and reactance can be represented as shown enclosed in dotted lines, for capacitive reactance. If the reactance is inductive, a coil is substituted for the capacitance C.

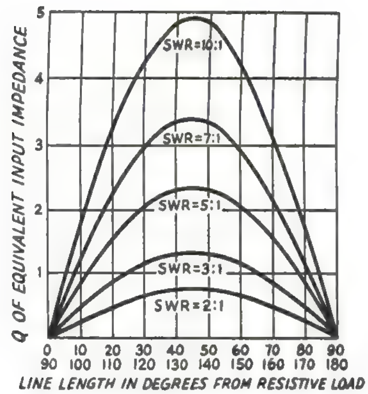


Fig. 3-33—Q of line input impedance as a function of line length and s.w.r. These curves repeat every 90 degrees of electrical line length. Q is defined as the ratio of reactance to resistance for the equivalent series circuit, and as the ratio of resistance to reactance for the equivalent parallel circuit.

reactance is low compared with the resistance in the parallel equivalent circuit, or when the reactance is high compared with the resistance in the series equivalent circuit. The very considerable detuning of the circuit that results is often accompanied by an increase in Q, sometimes to values that lead to excessively-high circulating currents in the circuit. This causes the efficiency to suffer. (Ordinarily the power loss in matching circuits of this type is inconsequential, if the Q is below 10 and a gool coil is used.) An unfavorable ratio of reactance to resistance in the input impedance of the line can exist if the s.w.r. is high and the line length is near an odd multiple of one-eighth wavelength (45 degrees).

Q of Line Input Impedance

The ratio between reactance and resistance in the equivalent input circuit—that is, the Q of the line input impedance—is shown in Fig. 3-33 as a function of line length and s.w.r. There is no specific value of this Q of which it can be said that lower values are satisfactory while higher values are not. In part, the maximum tolerable value depends on the tuning range available in the matching circuit. If the tuning range is restricted (as it will be if the variable capacitor has relatively low maximum capacitance), compensating for the line input reactance by absorbing it in the matching circuit—that is, by retuning C_1 in Fig. 3-32—may not be possible. Also, if the Q of the matching circuit is low the effect of the line input reactance will be greater than it will when the matching-circuit Q is high.

As stated earlier, the optimum matching-circuit design is one in which the Q is low. The lowest usable value of Q is determined by the coefficient of coupling k, between L_1 and L_2, Figs. 3-31 and 3-32. With practicable coil construction, values of k range between about 0.3 and 0.6, and the corresponding usable Q values in the matching circuit, assuming that the optimum inductance is used for L_2, range from 10 to 3, approximately. With a matching-circuit Q of 4 or 5, a tolerable value of the line input-impedance Q is approximately 2. It is to be expected, therefore, that a matching circuit using conventional constants and components for the frequency involved will be capable of handing all line lengths if the standing-wave ratio is 4 to 1 or less.

As shown by Fig. 3-33, difficulty may be expected at certain line lengths with higher standing-wave ratios, since the input-impedance Q is greater than 2 over a considerable portion of each 90-degree section.

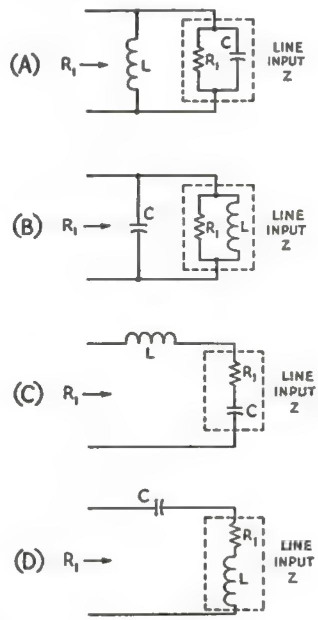


Fig. 3-34—Compensating for reactance present in the line input impedance.

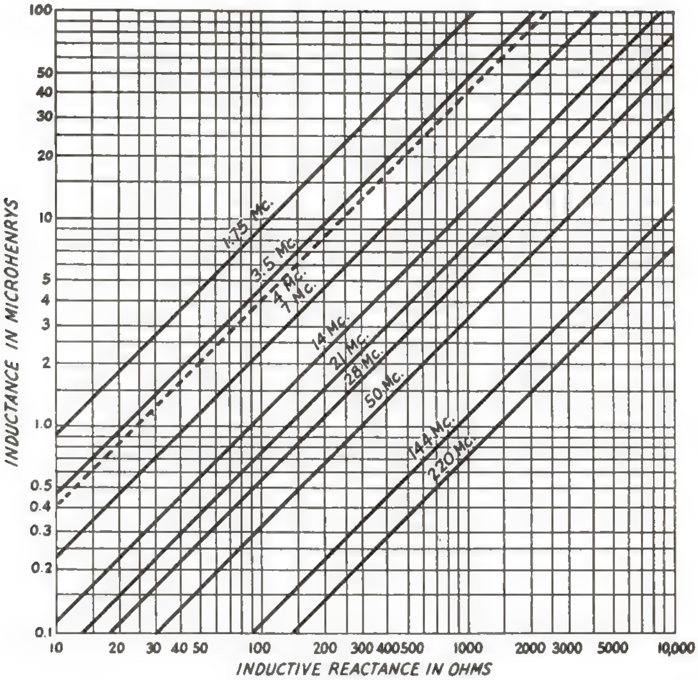


Fig. 3-35—Reactance vs. inductance for the various amateur bands. Values given are for the middle of each band, but will be sufficiently close for most calculations at any frequency within a band.

Compensating for Input Reactance

When the reactance/resistance ratio in the line input impedance is unfavorable it is advisable to take special steps to compensate for it. This can be done as shown in Fig. 3-34. Compensation consists in supplying external reactance of the same numerical value as the line reactance, but of the opposite kind. Thus in A, where the line input impedance is represented by resistance and capacitance in parallel, an inductance L having the same numerical value of reactance as C can be connected across the line terminals to "cancel out" the line reactance. (This is actually the same thing as tuning the line to resonance at the operating frequency.) Since the parallel combination of L and C is an extremely high resistance at resonance, the input impedance of the line becomes a pure resistance having essentially the same resistance as R_1 alone.

The case of an inductive line is shown at B. In this case the external reactance required is capacitive, of the same numerical value as the reactance of L.

Where the series equivalent of the line input impedance is used the external reactance is connected in series, as shown at C and D in Fig. 3-34.

In general, these methods need not be used unless C_1 in the matching circuit (Fig. 3-32) does not have sufficient range of adjustment to provide compensation for the line reactance as described earlier, or when such a large readjustment is required that the matching-circuit Q becomes undesirably high. The latter condition usually is accompanied by heating of L_1.

MATCHING-CIRCUIT CONSTANTS

The circuit of Fig. 3-31B and 3-32B will take care of nearly all coupling problems. The L/C ratio in the tunel circuit L_1C_1 is not critical, since the Q may be adjusted by the choice of positions for the line taps. However, unusually large values of C_1 for the frequency in use should be avoided since the coil losses tend to be higher in a high-C circuit. A reactance of approximately 500 ohms at the center of a band is a satisfactory value for L_1. The same value is required at C_1 to resonate the circuit. From Figs. 3-35 and 3-36 the corresponding values of inductance and capacitance are, approximately:

Band	*Inductance*	*Capacitance*
3.5 Mc.	22 μh.	90 pf.
7 Mc.	12 μh.	45 pf.
14 Mc.	6 μh.	23 pf.
21 Mc.	4 μh.	15 pf.
28 Mc.	2.9 μh.	12 pf.

The construction of the coils should be such that taps may be made to each turn.

As stated earlier, the optimum value for L_2 is the inductance that results in a reactance equal to the Z_0 of the coaxial line. For the various bands, the optimum inductance values for L_2 are as follows:

Band	*50-ohm line*	*75-ohm line*
3.5 Mc.	2.2 μh.	3.2 μh.
7 Mc.	1.2 μh.	1.7 μh.
14 Mc.	0.6 μh.	0.85 μh.
21 Mc.	0.4 μh.	0.56 μh.
28 Mc.	0.29 μh.	0.4 μh.

Methods for Variable Coupling

The coupling between L_1 and L_2 preferably should be adjustable. If the coupling is fixed, the placement of the taps on L_1 for proper matching becomes rather critical. The additional matching adjustment afforded by adjustable coupling between the coils facilitates the matching procedure considerably. L_2 should be coupled to the center of L_1 for the sake of maintaining balance, since the circuit is used with balanced lines.

If adjustable inductive coupling is not feasible for mechanical reasons, an alternative is to use a variable capacitor in series with L_2. This is shown in Fig. 3-37. Varying C_2 changes the total

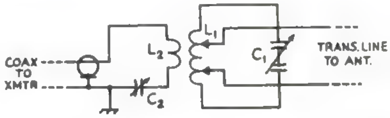


Fig. 3-37—Using a variable capacitance, C_2, as an alternative to variable mutual inductance between L_1 and L_2.

reactance of the circuit formed by L_2C_2, with much the same effect as varying the actual mutual inductance between L_1 and L_2. In this circuit L_2 can have the same optimum value of reactance recommended above, although it may be larger than the optimum if desired. The capacitance of C_2 should be such as to resonate with L_2 at the lowest frequency in the band. This calls for a fairly large value of capacitance at

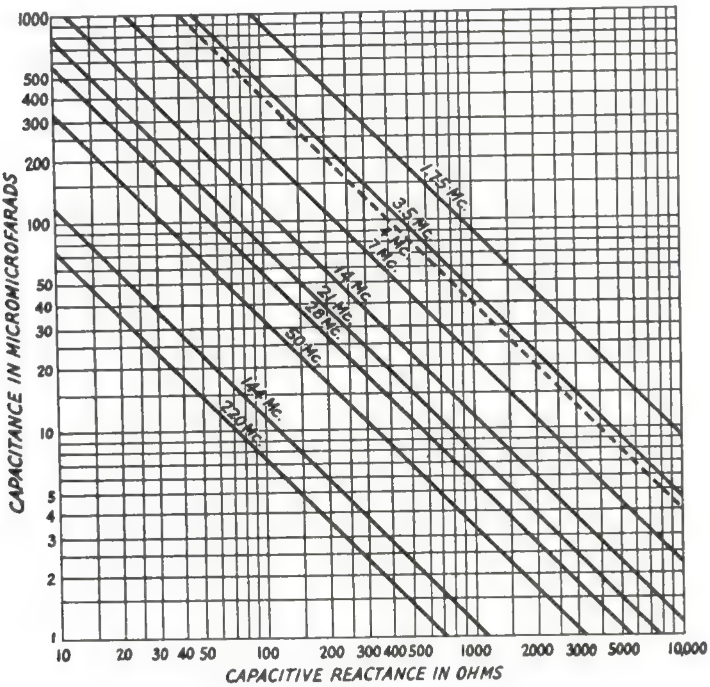


Fig. 3-36—Reactance vs. capacitance in μμf. (pf.) for the various amateur bands. Values given are for the middle of each band, but will be sufficiently close for most calculations at any frequency within a band.

low frequencies (about 1000 pf. at 3.5 Mc. for 50-ohm line) if the reactance of L_2 is equal to the line Z_0. To utilize a capacitor of more convenient size—maximum capacitance of perhaps 250-300 pf.—a value of inductance may be used for L_2 that will resonate at the lowest frequency with the maximum capacitance available.

On the higher-frequency bands the problem of variable capacitors does not arise since a reactance of 50 to 75 ohms is within the range of conventional components.

Circuit Balance

Fig. 3-37 shows C_1 as a balanced or split-stator capacitor. This type of capacitor is desirable in a practical matching circuit to be used with a balanced line, since the two sections are symmetrical while in the ordinary single-section capacitor there is more capacitance to ground (or metal objects, such as a chassis, in the vicinity) from the frame and rotor assembly than from the stator assembly. The rotor assembly of the balanced capacitor may be grounded, if desired. Alternatively, it may be left "floating" and the center of L_1 may be grounded; or both may "float." Which method to use depends on considerations discussed later in connection with antenna currents on transmission lines.

ADJUSTMENT OF INDUCTIVELY-COUPLED MATCHING CIRCUITS

Adjustment of the matching circuit consists in finding the proper settings for the taps on L_1, the proper setting of C_1, and the adjustment of the coupling between L_1 and L_2 so that the line input impedance is transformed into a 50- or 75-ohm load to match the Z_0 of the particular type of coaxial cable used between the transmitter and matching circuit. The surest and by far the simplest way to arrive at the correct adjustments is to use an s.w.r. bridge designed for the type of coax used.

The setup using the s.w.r. bridge is shown in Fig. 3-38. Adjustments may be made at any convenient power level within the power-handling capabilities of the type of bridge used. Adjust the transmitter output for a "forward" reading on the bridge indicator well up on the indicator scale. Then switch to read reflected power (or voltage, depending on the instrument calibration).

Adjustment Procedure

To adjust the circuit, set the taps at trial positions equidistant from the center of the coil, then vary C_1 for minimum reading on the bridge indicator. Next adjust C_2 for the lowest possible reading, then touch up C_1 again with the same object. Continue until the indicator reads zero or as close to it as possible. If a good null is not obtained, try the taps at another position and go through the procedure again. In most cases, it will not be necessary to try many tap positions; in fact, it is usually found that the tap positions are not at all critical. If C_2 is not used but the coupling between L_1 and L_2 can be varied, the coupling adjustment takes the place of varying C_2 in the above procedure. If neither C_2 nor adjustable coupling between the coils is used, the tap positions become rather critical.

During the course of adjustment, switch the bridge back to "forward" occasionally to make sure that this reading is staying well up on the scale. It sometimes happens that an adjustment which apparently reduces the reflected reading to zero or near zero is simply detuning the circuit and the forward reading becomes quite low. The object is to get a zero reflected reading at matching-circuit settings that also give a high forward reading.

After the initial adjustment of the circuit, try moving the tap positions out toward the ends of L_1 until it is just possible to obtain a match by means of C_1 and C_2 at all frequencies within the band. This will result in the lowest possible operating Q and thus minimize the necessity for readjustment of the circuit when shifting frequency.

With low Q, circuits of this type will work over an entire band without readjustment if the load is constant over the same frequency range. The load seldom stays constant, however, since the input impedance of the line changes with frequency with most antennas. Readjustment becomes necessary whenever the input impedance changes enough to result in poor operation. Evidence of this is either the inability to adjust the transmitter output circuit for proper loading of the final amplifier, or such a high s.w.r. in the coax that it shows signs of heating from the power lost in it.

Measurement of Line Input Current

The r.f. ammeters shown in Fig. 3-38 are not essential to the adjustment procedure but they, or some other form of output indicator, are useful accessories. In most cases the circuit adjustments that lead to a match as shown by the s.w.r. bridge will also result in the most efficient power transfer to the transmission line. However, it is possible, although not probable, that a good match will be accompanied by excessive loss in the circuit formed by L_1C_1. This is unlikely to happen if the taps on L_1 are kept as far apart as

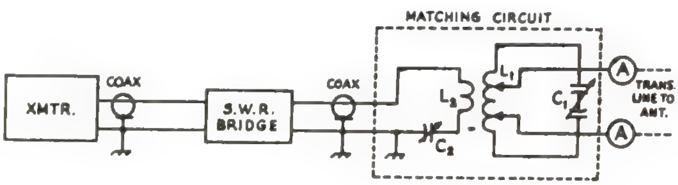


Fig. 3-38—Adjustment setup using s.w.r. bridge.

possible, since spreading the taps keeps the Q low. If the settings of C_1 and C_2 are highly critical and/or it is impossible to obtain a match unless the taps on L_1 are close together (considering the size of the coil) the use of additional reactance compensation as described earlier is indicated.

R.f. ammeters are useful for showing the comparative output obtained with various tap settings, and also for showing the improvement in output resulting from the use of reactance compensation when it seems to be required. Providing no basic circuit changes (such as grounding or ungrounding some part of the matching circuit) are made during such comparisons, the current shown by the ammeters will increase whenever the power put into the line is increased, and so the highest reading indicates the greatest transfer efficiency, assuming that the power input to the transmitter is kept constant.

If the line Z_0 is matched by the antenna the current can be used to determine the actual power input to the line. The power at the input terminals is then equal to I^2Z_0, where I is the current and Z_0 is the characteristic impedance of the line. If there are standing waves on the line this relationship does not hold. In such a case the current that will flow into the line is determined by the line length, s.w.r., and whether the antenna impedance is higher or lower than the line impedance. Fig. 3-19 shows how the maximum current to be expected will vary with the standing-wave ratio. This information can be used in selecting the proper ammeter range.

For example, assume that a 600-ohm open-wire line is to be used and that the power is 500 watts. The antenna is to be an all-band affair and the highest s.w.r. to be expected is about 10 to 1. With perfect matching, the current would be $\sqrt{500/600} = 0.91$ amp. From Fig. 3-19, the current at a loop will be 3.16 times the matched current when the s.w.r. is 10, so the maximum current to be expected under any conditions is $3.16 \times 0.91 = 2.9$ amp. An instrument having a range of 0-3 or 0-4 amperes would be suitable. The minimum current to be expected would be the maximum divided by the s.w.r., or $2.9/10 = 0.29$ amp.

Two ammeters, one in each line conductor, are shown in Fig. 3-38. The use of two instruments gives a check on the line balance, since the currents should be the same. However, a single meter can be switched from one conductor to the other. If only one instrument is used it is preferably left out of the circuit except when adjustments are being made, since it will add capacitance to the side in which it is inserted and thus cause some unbalance. This is particularly important when the instrument is mounted on a metal panel.

Since the resistive component of the input impedance of a line operating with an appreciable s.w.r. is seldom known accurately, the r.f. current is of little value as a check on power input to such a line. However, it shows in a relative way the efficiency of the system as a whole. The set of coupling adjustments that results in the largest line current with the least final-amplifier plate current is the one that delivers the greatest power to the antenna with the lowest plate power input. Also, in case an amplifier is not opeated properly—for example, the grid excitation may be low—the r.f. meter will show when maximum power output has been obtained. In such cases maximum output, as shown by the largest obtainable r.f. meter reading, may occur at a value of plate current less than the ratings of the amplifier tubes.

For adjustment purposes, it is possible to substitute small flashlight lamps, shunted across a few inches of the line wires, for the r.f. ammeters. Their relative brightness shows when the current increases or decreases. They have the advantage of cheapness and such small physical size that they do not unbalance the circuit.

Adjustment without S.W.R. Indicator

An s.w.r. indicator (particularly one built primarily for use in impedance-matching adjustments rather than for accurate measurement of s.w.r.) can be such an inexpensive and easily-made instrument that there is no real necessity for trying to get along without one. Without the bridge it is seldom possible to arrive at the best adjustment for the matching circuit except by making a series of trials during which the power input to the line is compared with the plate input to the final amplifier. This is often a laborious process, and requires using r.f. ammeters or an equivalent means for indicating current in the tansmission line.

The adjustment procedure is one of cut-and-try, using various positions of the line taps on L_1, and, for each set of tap positions, varying C_1, C_2 and the transmitter output tuning and coupling to give a desired value of final-amplifier plate current (the plate tank being resonated each time). When the desired loading has been obtained, the r.f. output as shown by the line current should be recorded, and then the same procedure followed again with a new set of tap positions on L_1. Eventually a set of adjustments giving the highest r.f. current into the line, for a fixed plate input, will be found. This represents the best adjustment of the matching circuit.

Coaxial-Line Feed

As mentioned earlier, a matching circuit should not be necessary when coaxial line is used to feed the antenna, since the s.w.r. on such lines already should be low enough to permit satisfactory adjustment of the coupling by the normal transmitter controls. However, there are cases where the additional frequency selectivity provided by the matching circuit is desirable. For example, a coax-fed multiband antenna system will not discriminate against transmitter harmonics, since the system is designed to accept harmonically related amateur frequencies without individual tuning adjustments.

With a slight rearrangement, the matching circuit can readily be adapted to coaxial lines. This is shown in Fig. 3-39. In A, the circuit itself is identical with that in Fig. 3-38, but one tap (that connected to the outside of the coax line) is connected to the center of the coil. The center conductor of the line is tapped on the coil on one side of center (either side will do). In B a single-ended circuit is substituted, since a balanced circuit is not needed with coax. The value of L_1 should be the same as in the previous circuit; likewise for C_1 in A. In B, C_1 may be a single-ended capacitor having a value half that of the per-section value in A.

The circuit is adjusted in the same way as in the case of balanced line.

RATINGS OF COMPONENTS

The currents and voltages in the matching circuit can vary over a wide range, depending principally on the operating Q of the circuit. Assuming a maximum Q of 10 and the inductance and capacitance values recommended earlier, the maximum current to be expected in the circuit formed by L_1 and C_1, Figs. 3-37 or

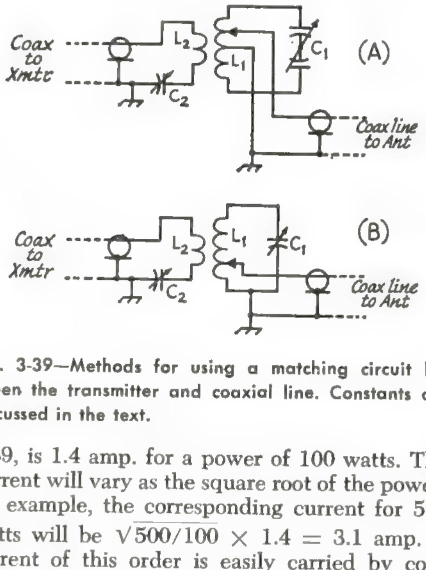


Fig. 3-39—Methods for using a matching circuit between the transmitter and coaxial line. Constants are discussed in the text.

3-39, is 1.4 amp. for a power of 100 watts. The current will vary as the square root of the power; for example, the corresponding current for 500 watts will be $\sqrt{500/100} \times 1.4 = 3.1$ amp. A current of this order is easily carried by coils wound with No. 14 wire, and No. 12 is satisfactory for any power likely to be developed by an amateur transmitter up to a kilowatt input. For 100 watts or less No. 16 or 18 is sufficiently large. If the Q can be kept low the current will be smaller for a given power level. The recommended wire sizes are adequate for key-down c.w. or amplitude modulated phone; in s.s.b. transmission the average current is much smaller for the equivalent peak-envelope power level so the heating effect is less. Under average conditions, a key-down r.f. power level of 100 watts should cause about the same heating as 500 watts p.e.p. output in s.s.b.

Under the same assumptions with respect to circuit conditions the r.f. peak voltage that the capacitor must stand is 1000 volts (700 volts r.m.s.) for a power of 100 watts without modulation. With 100 per cent amplitude modulation the peak voltage will be double this, or 2000 volts. The voltage also varies as the square root of the power level, so at the maximum output of about 750 watts from a one-kilowatt input c.w. transmitter the voltage can be as high as 2750 volts, or 5500 volts with 100 per cent amplitude modulation. A 2-kw. p.e.p. input single-sideband transmitter, which normally will have a p.e.p. r.f. output of 1200 to 1400 watts, would develop about 3750 volts at the r.f. peaks. These figures will be reduced if the Q is kept below 10.

If C_1 is a balanced capacitor, as in Fig. 3-37, or consists of two capacitors in the series-tuned circuit, Fig. 3-39, the voltage that each section must stand is one-half the total.

Since the purpose of the matching circuit is to tansform the actual line input impedance into 50 to 75 ohms, the current carried by L_2 in the matching circuit will be the current that will flow in a matched line; that is, the current is equal to the square root of the power divided by Z_0. For 50-ohm line it is 1.4 amp. and for 75-ohm line it is 1.15 amp., at the 100-watt level. The current in L_2 under the condition of proper matching is thus of the same order as the current in L_1 under the extreme condition of a Q of 10, so the wire sizes mentioned above for various power levels will be satisfactory.

The voltage rating required for C_2 will depend on the actual capacitance used. The peak voltage across this capacitor is equal to the peak current multiplied by the reactance of the capacitor. If this reactance is approximately equal to the line Z_0, the voltage will be quite low (about 100 volts for 100 watts in 50-ohm line, for example) and receiving-type capacitors will be adequate even for rather high power. However, if the reactance is several times the line Z_0 (capacitance comparatively small) the required voltage rating increases in proportion to the reactance (inverse proportion to the capacitance).

BALUN COILS

In Fig. 3-40, L_1 is a bifilar winding which, when considered as a pair of parallel conductors, is equivalent to a transmission line. To a voltage applied between the two terminals at one end of the winding the inductance of the winding considered as a plain coil is unimportant, since the currents in the two conductors will be equal and opposite and there is substantially complete cancellation of the external field just as in a normal transmission line. The parallel conductors have a characteristic impedance dependent on their diameter and spacing. L_2 is an identical winding.

If the two windings are connected as shown at A, the two transmission lines are in series at the right-hand end and in parallel at the left-hand end. If each has a characteristic impedance equal to half the input resistance of the balanced transmission line to the antenna, their Z_0's will be

matched by the line input resistance and the resistance looking into the left-hand end of each will be equal to its own Z_0. Since they are in parallel at the left-hand end, the total resistance looking into the windings toward the antenna is equal to half the Z_0 of either line, or one-fourth R_1.

This arrangement therefore acts as an impedance transformer having a fixed ratio of 4 to 1, balanced on the high-impedance side and unbalanced on the low-impedance side. Since a ratio of 4 to 1 happens also to be the ratio between matched 300-ohm twin line and 75-ohm coaxial line, such a balancing circuit or **balun** is useful for matching a flat 300-ohm line to 75-ohm coax.

For parallel or "antenna" currents on the transmission line to the antenna (see later section in this chapter) the two pairs of windings act as normal inductances, since such currents are in phase in the two line wires. They thus tend to choke off parallel currents. This choking action also is essential or keep the right-hand end of the lower conductor of L_2 properly above ground for balanced output. Thus the coils have to be sufficiently large to give good isolation between the balanced and unbalanced ends at the lowest frequency to be used. The upper frequency limit is that at which the winding, considered as an inductance rather than a transmission line, begins to show distributed-capacitance effects. In practice, a single set of coils can be designed to work over the 3.5- to 30-Mc. range. Design is complicated because there is mutual coupling between turns, which modifies the characteristic impedance. However, suitable units are available commercially.

The principles on which balun coils operate should make it obvious that the s.w.r. on the transmission line to the antenna must be close to 1 to 1. If it is not, the input impedance of each bifilar winding will depend on its electrical characteristics and the input impedance of the main transmission line; in other words, the impedance will vary just as it does with any transmission line and the transformation ratio likewise will vary over wide limits. Hence it is essential that the antenna match the 300-ohm twin line at all frequencies at which the balun assembly will be used.

Balun coils are convenient, when the above condition can be met, since they require no adjustment. However, spurious components in the transmitter's output (such as harmonics and "subharmonic" leakage from buffer and multiplier stages) are transferred to the antenna through them as readily as the desired output frequency. The matching circuits described earlier add considerable selectivity to the system and thus tend to suppress such undesired radiations.

"TUNED" AND "UNTUNED" LINES

In the past, transmission lines frequently have been classified as "tuned" or "untuned," depending on whether or not the line had to be cut to a certain length in order to have a substantially-resistive input impedance. As shown earlier, when the s.w.r. is high the input impedance is resistive (or mostly so) only at line lengths that are a multiple of a quarter wavelength, assuming the load represented by the antenna is itself a substantially pure resistance. (If it is not, the resistive-input points correspond to the current and voltage loops, the first of which may occur at any distance up to a quarter wave length from the antenna, depending on the antenna impedance. After the first loop, the remaining ones are at quarter-wavelength intervals.)

Such a classification tends to be arbitrary, since there is no well-defined value of s.w.r. below which a line may be "untuned" and above which it must be "tuned." It is possible to couple power into *any* length of line, regardless of the standing-wave ratio, if the principles already outlined are followed. If the s.w.r. is high, special methods of reactance compensation may be required if the line length is unfavorable, as explained earlier, but it is not *necessary* to cut the line to a particular length (which may be an inconvenient length for installation) in order to put power into it.

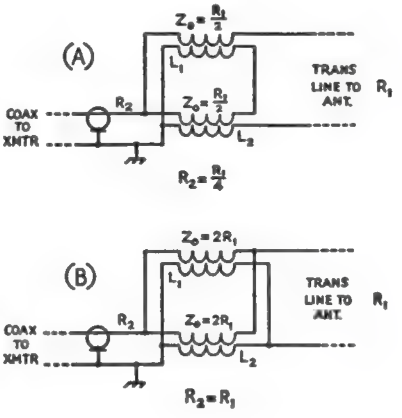


Fig. 3-40—Balun coils as impedance-matching device for coupling between balanced and unbalanced lines. A properly-designed set of coils will wor kover the 3.5-30-Mc. range without adjustment, but only the fixed impedance ratios shown are available and the input impedance of the transmission line must be a pure resistance that will match the Z_0 of the coils.

Coupling The Line To The Antenna

Throughout the discussion of transmission-line principles in the first part of this chapter the operation of the line has been described in terms of an abstract "load." This load had the electrical properties of resistance and, sometimes, reactance. It did not, however, have any physical attributes that associated it with a particular electrical device. That is, it could be anything at all that exhibits electrical resistance and/or reactance. The fact is that so far as the line is concerned, it does not matter what the load *is*, just so long as it will accept power.

Many amateurs make the mistake of confusing transmission lines with antennas, believing that because two identical antennas have different kinds of lines feeding them, or the same kind of line with different methods of coupling to the antenna, the "antennas" are different. There may be practical reasons why one system (including antenna, transmission line, and coupling method) may be preferred over another in a particular application. But to the transmission line an antenna is just a load that terminates it, and the important thing is what that load looks like to the line in terms of resistance and reactance.

Any kind of transmission line can be used with *any* kind of antenna, if the proper measures are taken to couple the two together.

Frequency Range and S.W.R.

Probably the principal factor that determines the way a transmission line is operated is the frequency range over which the antenna is to work. Very few types of antennas will present essentially the same load impedance to the line on harmonically-related frequencies. As a result, the builder often is faced with choosing between an antenna system that will permit operating the transmission line with a low standing-wave ratio, but is confined to one operating frequency or a narrow band of frequencies, and a system that will permit operation in several harmonically-related bands but with a large s.w.r. on the line. (There are "multiband" systems which, in principle, make one antenna act as though it were a half-wave dipole on each of several amateur bands, by using "trap" circuits or multiple wires. Data on these is given in later chapters. Such an antenna can be assumed to be equivalent to a resonant (half-wave) dipole on each of the bands for which it is designed, and may be fed through a line, coaxial or otherwise, that has a Z_0 matching the antenna impedance, as described later in this chapter for simple dipoles.)

Methods of coupling the line to the antenna therefore divide, from a practical standpoint, into two classes. In the first, operation on several amateur bands is the prime consideration and the standing-wave ratio is secondary. The s.w.r. is normally rather large and the input impedance of the line depends on the line length and the operating frequency.

In the second class, a conscious attempt is made, when necessary, to transform the antenna impedance to a value that matches the characteristic impedance of the line. When this is done the line operates with a very low standing-wave ratio and its input impedance is essentially a pure resistance, regardless of the line length. A transmission line can be considered to be "flat," within practical limits, if the s.w.r. is not more than about 1.5 to 1.

Losses

A principal reason for matching the antenna to the line impedance is that a flat line operates with the least power loss. While it is always desirable to reduce losses and thus increase efficiency, the effect of standing waves in this connection can be overemphasized. This is particularly true at the lower amateur frequencies. where the inherent loss in most types of lines is quite low even for runs that, in the average amateur installation, are rather long.

For example, 100 feet of 300-ohm receiving-type Twin-Lead has a loss of only 0.18 db. at 3.5 Mc., as shown by Table 3-I. Even with an s.w.r. as high as 10 to 1 the additional loss caused by standing waves is less than 0.7 db., from Fig. 3-18. Since 1 db. represents the minimum detectable change in signal strength, it does not matter from this standpoint whether the line is flat or not. But at 144 Mc. the loss in the same length of line perfectly matched is 2.8 db., and an s.w.r. of 10 to 1 would mean an *additional* loss of 4 db. At the higher fre-

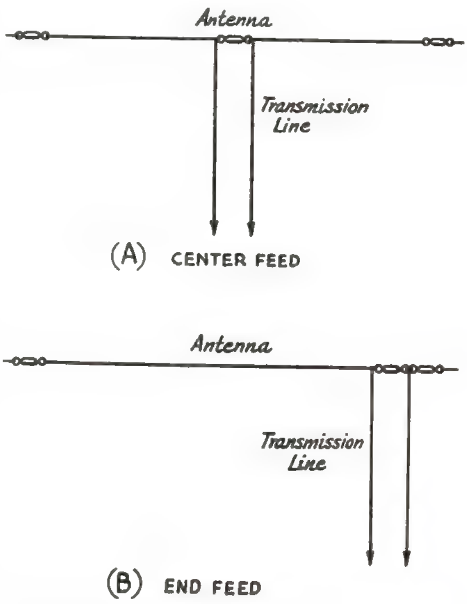


Fig. 3-41—Center and end feed as used in simple antenna systems.

quency, then, it is worth while to match the antenna and line as closely as possible.

Power Limitations

Another reason for matching is that certain types of lines, particularly those with solid dielectric, have definite voltage and current limitations. At the lower frequencies this is a far more compelling reason than power loss for at least approximate matching. Where the voltage and current must not exceed definite maximum values, the amount of power that the line can handle is inversely proportional to the standing-wave ratio. If the safe rating on the 300-ohm line in the example above is 500 watts when perfectly matched, the line can handle only 50 watts with equal safety when the s.w.r. is 10 to 1. Thus, despite the fact that the line losses are low enough to make no appreciable difference in the signal strength the high s.w.r. could be tolerated only with low-power transmitters.

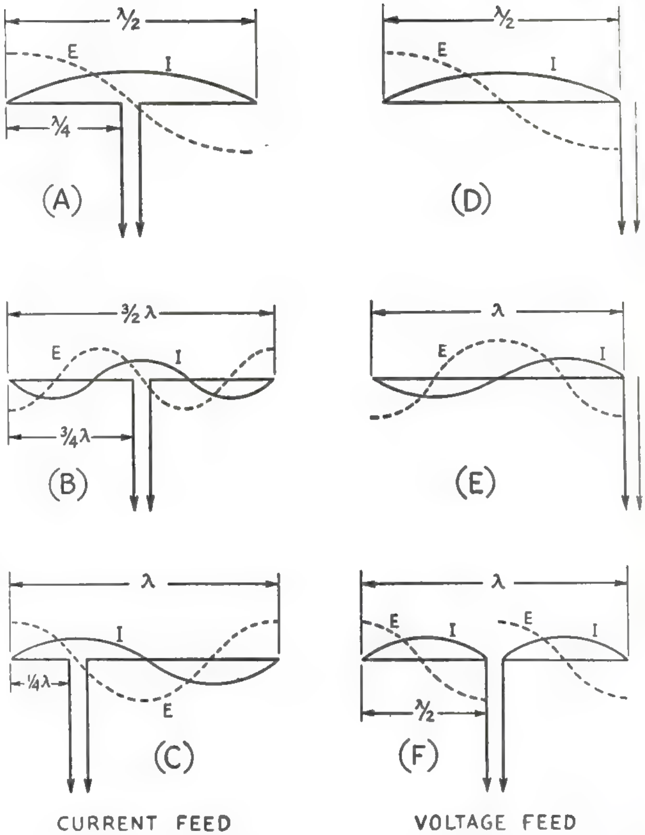


Fig. 3-42—Current and voltage feed, in antennas operated at the fundamental frequency, 2 times the fundamental and 3 times the fundamental. The current and voltage distribution on the antenna are identical with both methods only at the fundamental frequency.

Line Radiation

Aside from power considerations, there is a more-or-less common belief that a flat line "does not radiate" while one with a high s.w.r. does radiate.

This impression is quite unjustified. It is true that the radiation from a parallel-conductor line increases with the current in the line, and that the effective line current increases with the s.w.r. However, the loss by radiation from a properly balanced line is so small (and is, furthermore, independent of the line length) that multiplying it several times still does not bring it out of the "negligible" classification.

Whenever a line radiates it is because of faulty installation (resulting in unbalance with parallel-conductor lines) or "antenna currents" on the line. Radiation from the latter cause can take place from either resonant or nonresonant lines, parallel-conductor or coaxial.

UNMATCHED SYSTEMS

In many multiband systems or simple antennas where no attempt is made to match the antenna impedance to the characteristic impedance of the line, the customary practice is to connect the line either to the center of the antenna (center feed) as indicated in Fig. 3-41A, or to one end (end feed) as shown in Fig. 3-41B.

Because the line operates at a rather high standing-wave ratio, the best type to use is the open-wire line. Solid Twin-Lead of the 300-ohm receiving variety can also be used, but the power limitations discussed in the preceding section should be kept in mind. Although the manufacturers have placed no power rating on receiving-type 300-ohm line it seems reasonable to make the assumption, based on the conductor size, that a current of 2 amp. can readily be carried by a line installed so that there is free air circulation about it. This corresponds to a power of 1200 watts in a matched 300-ohm line. When these are standing waves, the safe power can be found by dividing 1200 by the s.w.r. In a center-fed half-wave antenna as in Fig. 3-41A, the s.w.r. should not exceed about 5 to 1 (at the fundamental frequency), so receiving type 300-ohm Twin-Lead would appear to be safe for power outputs up to 250 watts or so.

Since there is little point in using a mismatched line to feed an antenna that is to operate on one amateur band only, the discussion to follow will be based on the assumption that the antenna is to be operated on its harmonics for multiband work.

"Current" and "Voltage" Feed

Usual practice is to connect the transmission line to the antenna at a point where either a current or voltage loop occurs. If the feed point is at a current loop the antenna is said to be current fed; if at a voltage loop the antenna is voltage fed.

These terms should not be confused with center feed and end feed, because they do not necessarily have corresponding meanings. There is always a voltage loop at the end of a resonant antenna, no matter what the number of half wavelengths, so a resonant end-fed antenna is always voltage fed. This is illustrated at D and E in Fig. 3-42 for end-fed antennas a half wavelength long (antenna fundamental frequency) and one wavelength long (second harmonic). It would continue to be true for an end-fed antenna operated on any harmonic. However, Fig. 3-42F shows voltage feed at the *center* of the antenna; in this case the antenna has a total length of two half wavelengths each of which is voltage fed. Voltage feed is determined not by the physical position of the transmission line on the antenna, but by the fact that a voltage loop occurs on the antenna at the feed point. Since voltage loops always occur at integral multiples of a half wavelength from either end of a resonant antenna, feeding the antenna at any half-wavelength point constitutes voltage feed.

Typical cases of current feed are shown at A, B and C in Fig. 3-42. The feed point is at a current loop, which always occurs at the midpoint of a half-wave section of the antenna. In order to feed at a current loop the transmission line must be connected at a point that is an odd multiple of one-quarter wave-length from either end of the resonant antenna. A center-fed antenna is also current fed *only* when the antenna length is an *odd* multiple of one-half wavelength. Thus the antenna in Fig. 3-42B is both center fed and current fed since it is three half wavelengths long. It would also be center fed and current fed when five, seven, etc., half wavelengths long.

To current feed a one-wavelength antenna, or any resonant antenna having a length that is an *even* multiple of one-half wavelength, it is necessary to shift the feed point from the center of the antenna (where a voltage loop always occurs in such a case) to the middle of one of the half-wave sections. This is indicated in Fig. 3-42C in the case of a one-wavelength antenna; current feed can be used if the line is connected to the antenna at a point ¼ wavelength from either end.

Operation on Harmonics

In the usual case of an antenna operated on several bands, the point at which the transmission line is attached is of course fixed. The antenna length is such that it is resonant at some frequency in the lowest-frequency band to be used and the transmission line is connected either to the center or the end. The current and voltage distribution along antennas fed at both points is shown in Fig. 3-43. With end feed, A to F inclusive, there is always a voltage loop at the feed point. Also, the current distribution is such that in every case the antenna operates as a true harmonic radiator of the type described in Chapter Two.

With center feed, the feed point is always at a current loop on the fundamental frequency and all *odd* multiples of the fundamental. In these cases the current and voltage distribution

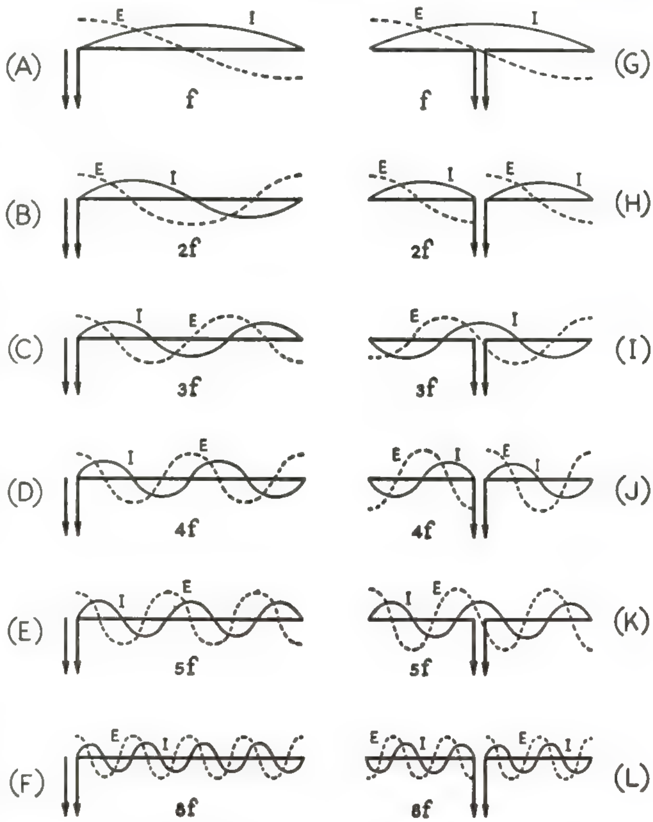


Fig. 3-43—Current and voltage distribution at the fundamental frequency and various multiples, with both end feed and center feed. The distributions are the same with both types of feed only when the frequency is an odd multiple of the fundamental.

are identical with the distribution on an end-fed antenna. This can be seen by comparing A and G, C and I, and E and K, Fig. 3-43. (In I, the phase is reversed as compared with C, but this is merely for convenience in drawing; the actual phases of the currents in each half-wave section reverse each half cycle so it does not matter whether the current curve is drawn above or below the line, so long as the *relative* phases are properly shown in the same antenna.) On odd multiples of the fundamental frequency, therefore, the antenna operates as a true harmonic antenna.

On *even* multiples of the fundamental frequency the feed point with center feed is always at a voltage loop. This is shown at H, J and L in Fig. 3-43. Comparing B and H, it can be seen that the current distribution is different with center feed than with end feed. With center feed the currents in both half-wave sections of the antenna are in the same phase, but with end feed the current in one half-wave section is in reverse phase to the current in the other. This does not mean that one antenna is a better radiator than the other, but simply that the two will have different directional characteristics. The center-fed arrangement is commonly known as "two half-waves in phase," while the end-fed system is a "one-wavelength antenna" or "second-harmonic" antenna.

Similarly, the system at J has a different current and voltage distribution than the system at D, although both resonate at four times the fundamental frequency. A similar comparison can be made between F and L. The center-fed arrangement at J really consists of two one-wavelength antennas, while the arrangement at L has two 2-wavelength antennas. These have different directional characteristics than the 2-wavelength and 4-wavelength antennas (D and F) that resonate to the same multiple, respectively, of the fundamental frequency.

The reason for this difference between odd and even multiples of the fundamental frequency in the case of the center-fed antenna can be explained with the aid of Fig. 3-44. It will be recalled from Chapter Two that the direction of current flow reverses in each half wavelength of wire. Also, in any transmission line the currents in the two wires always must be equal and flowing in opposite directions at any point along the line. Starting from the end of the antenna, the current must be flowing in one direction throughout the first half-wavelength section, whether this section is entirely antenna or partly antenna and partly one wire of the transmission line. Thus, in A, Fig. 3-44, the current flows in the same direction from *P* to *Q*, since this is all the same conductor. However, one quarter wave is in the antenna and one in the transmission line. The current in the other line wire, starting from *R*, must flow in the opposite direction in order to balance the current in the first wire, as shown by the arrow. And since the distance from *R* to *S* is ½ wavelength, the current continues to flow in the same direction all the way to S. The currents in the two halves of the *antenna* are therefore flowing in the same direction. Furthermore, the current is maximum ¼ wavelength from the ends of the antenna, as

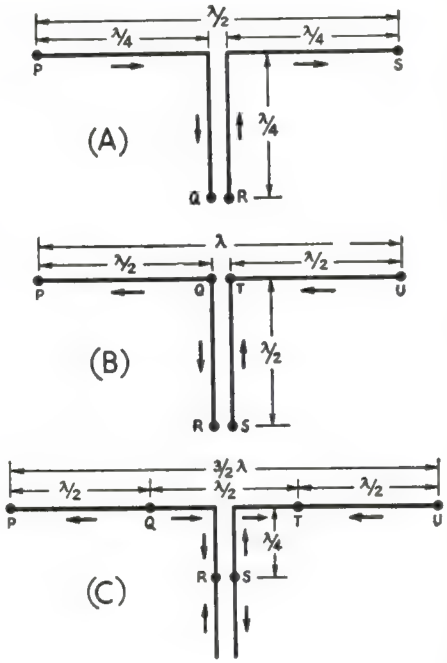


Fig. 3-44—Showing how the type of feed changes from current to voltage, with a center-fed antenna, on twice the fundamental frequency, and back to current feed on three times the fundamental. The same change occurs between all even and odd multiples.

previously explained, and so both the currents are maximum at the junction of the antenna and transmission line. This makes the current distribution along the antenna exactly the same as with end feed.

Fig. 3-44B shows the case where the overall length of the antenna is one wavelength, making a half wave on each side. A half wavelength along the transmission line also is shown. If we assume that the current is flowing downward in the line conductor from *Q* to *R*, it must be flowing upward from *S* to *T* if the line currents are to balance. However, the distance from *Q* to *P* is ½ wave length, and so the current in this section of the antenna must flow in the opposite direction to the current flowing in the section from *Q* to *R*. The current in section *PQ* is therefore flowing *away* from *Q*. Also, the current in section *TU* must be flowing in the opposite direction to the current in *ST*, and so is flowing *toward T*. The currents in the two half-wave sections of the antenna are therefore flowing in the same direction. That is, they are in the same phase.

With the above in mind, the direction of current flow in a 1½-wavelength antenna, Fig. 3-44C, should be easy to follow. The center half-wave section QT corresponds to the half-wave antenna in A. The currents in the end sections, PQ and TU simply flow in the opposite direction to the current in QT. Thus the currents are out of phase in alternate half-wave sections.

The shift in voltage distribution between odd and even multiples of the fundamental frequency can be demonstrated by a similar method, making allowance for the fact that the voltage is maximum where the current is minimum, and vice versa. On all even multiples of the fundamental frequency there is a current minimum at the junction of the line and antenna, with center feed, because there is an integral number of half wavelengths in each side of the antenna. The voltage is maximum at the junction in such a case and we have voltage feed. Where the multiple of the fundamental is odd, there is always a current maximum at the junction of the transmission line and antenna, as demonstrated by A and C in Fig. 3-44. At these points the voltage is minimum and we therefore have current feed.

"Zepp" or End Feed

In the early days of short-wave communication an antenna consisting of a half-wave dipole end-fed through a ¼-wavelength transmission line was developed as a trailing antenna for Zeppelin airships. In its utilization by amateurs, over the years, it has become popularly known as the "Zeppelin" or "Zepp" antenna. The term is now applied to practically any resonant antenna fed at the end by a two-wire transmission line.

The mechanism of end feed is perhaps somewhat difficult to visualize, since only one of the two wires of the transmission line is connected to the antenna while the other is simply left free. The difficulty lies in the natural tendency to think in terms of current flow in ordinary electrical circuits, where it is necessary to have a complete loop between both terminals of the power source before any current can flow at all. But as explained earlier, this limitation applies only to circuits in which the electromagnetic fields reach the most distant part of the circuit in a time interval that is negligible in comparison with the time of one cycle. When the circuit dimensions are comparable with the wavelength no such complete loop is necessary. The antenna itself is an example of an "open" circuit in which large currents can flow.

One way of looking at end feed is to consider the entire length of wire, including both antenna and feeder, as a single unit. For example, suppose we have a wire one wavelength long, as in Fig. 3-45A, fed at a current loop by a source of r.f. power. The current distribution will be as shown by the curves, with the assumed directions indicated by the arrows. If we now fold back the ¼- wavelength section to the left of the power source, as shown at B, the over-all current distribution will be similar, but the currents in the two wires of the folded section will be flowing in opposite directions. The amplitudes of the currents at any point along the folded-back portion will be equal in the two wires. The folded section therefore has become a ¼-wavelength transmission line, since the fields from the equal and opposite currents cancel. There

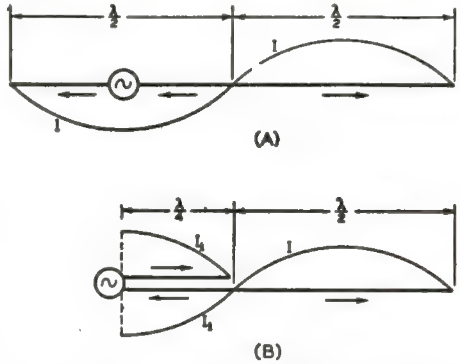


Fig. 3-45—Folded-antenna analogy of transmission line for an end-fed antenna.

is, however, nothing to prevent current from continuing to flow in the right-hand half-wavelength section, since there was current there before the left-hand section was folded.

This picture, although showing how power can flow from the transmission line to an antenna through end feed, lacks completeness. It does not take into account the fact that the current I_1 in the transmission line is greatly different from the current I in the antenna, A more basic viewpoint is the one already mentioned in Chapter Two: the current is caused by electromagnetic fields traveling along the wire and simply constitutes a measurable manifestation of those fields; the current does not cause the fields. From this standpoint the transmission-line conductors merely serve as "guides" for the fields so the electromagnetic energy will go where we want it to go. When the energy reaches the end of the transmission line it meets another guide, in the form of the antenna, and continues along it. However, the antenna is a different form of guide: it has a single conductor while the line has two; it has no provision for preventing radiation while the line is designed for that very purpose. This is simply another way of saying that the impedance of the antenna differs from that of the transmission line, so there will be reflection when the energy traveling along the line arrives at the antenna. We are then back on familiar ground, in that we have a transmission line terminated in an impedance different from its characteristic impedance.

Feeder Unbalance

With end feed, the currents in the two line wires do not balance exactly and there is there-

fore some radiation from the line. The reason for this is that the current at the end of the free wire is zero (neglecting a small charging current in the insulator at the end) while the current does not go to zero at the junction of the "active" line wire and the antenna. This is because not all the energy going into the antenna is reflected back from the far end, some being radiated; hence the incident and reflected currents cannot completely cancel at a node.

In addition to this unavoidable line radiation a further unbalance will occur if the antenna is not exactly resonant at the operating frequency. If the frequency is too high (antenna too long) the current node does not occur at the junction of the antenna and "live" feeder but moves out on the antenna. When the frequency is too low the node moves down the active feeder. Since the node on the free feeder has to occur at the end, either case is equivalent to shifting the position of the standing wave along one feeder wire but not the other. Representative cases are shown in Fig. 3-46. The farther off resonance the antenna is operated the greater the unbalance and the greater the line radiation. With center feed this unbalance does not occur, as is also shown in Fig. 3-46, because the system is symmetrical with respect to the line.

To avoid line radiation it is always best to feed the antenna at its center of symmetry. In the case of simple antennas for operation in several bands, this means that center feed should be used. End feed is required only when the antenna is operated on an even harmonic to obtain a desired directional characteristic, and then only when it must be used on more than one band. For single-band operation it is always possible to feed an even-harmonic antenna at a current loop in one of the half-wave sections nearest the center.

S.W.R. with Wire Antennas

When a line is connected to a single-wire antenna at a current loop the standing-wave ratio can be estimated with good-enough accuracy with the aid of the curve in Fig. 2-23. Although the actual value of the radiation resistance, as measured at a current loop, will vary with the height of the antenna above ground, the theoretical values given in Fig. 2-23 will at least serve to establish whether the s.w.r. will be high or low.

With center feed the line will connect to the antenna at a current loop on the fundamental frequency and all odd multiples, as shown by Fig. 3-43. At the fundamental frequency and usual antenna heights, the antenna resistance should lie between 50 and 100 ohms, so with a line having a characteristic impedance of 450 ohms the s.w.r. will be $Z_0/R_L = 450/50 = 9$ to 1 as one limit and $450/100 = 4.5$ to 1 as the other. On the third harmonic the theoretical resistance as given by Fig. 2-23 is 100 ohms, so the s.w.r. should be about 4.5 to 1. For 300-ohm line the s.w.r. can be expected to be between 3 and 6 on the antenna fundamental and about 3 to 1 on the third harmonic.

The impedances to be expected at voltage loops are less readily determined. Theoretical values are in the neighborhood of 5000 to 8000 ohms, depending on the antenna-conductor size and the number of half wavelengths along the wire. Such experimental figures as are available indicate a lower order of resistance, with measurements and estimates running fom 1000 to 5000 ohns. In any event, there will be some difference between end feed and center feed, since the current distribution on the antenna is different in these two cases at any given even multiple of the fundamental frequency. Also, the higher the multiple the lower the resistance at a voltage loop, so the s.w.r. can be expected to decrease when an antenna is operated at a high multiple of its fundamental frequency. Assuming 4000 ohms for a wire antenna two half waves long, the s.w.r. would be about 6 or 7 with a 600-ohm line and around 12 with a 300-ohm line. However, considerable variation is to be expected.

ANTENNA CURRENTS ON TRANSMISSION LINES

In any discussion of transmission-line operation it is always assumed that the two conductors carry equal and opposite currents throughout their length. This is an ideal condition that may or may not be realized in practice. In the

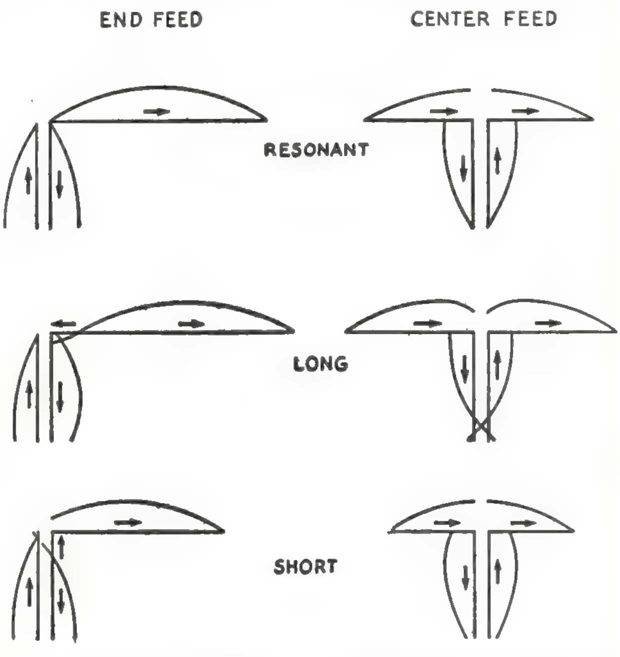


Fig. 3-46—Effect of departure from exact resonance in antenna with end-fed and center-fed systems.

average case the chances are rather good that the currents will *not* be balanced unless special precautions are taken. Whether the line is matched or not has little to do with the situation.

Consider the half-wave antenna shown in Fig. 3-47 and assume that it is somehow fed by a source of power at its center, and that the instantaneous direction of current flow is as indicated by the arrows. In the neighborhood of the antenna is a group of conductors disposed in various ways with respect to the antenna itself. All of these conductors are in the field of the antenna and are therefore coupled to it. Consequently, when current flows in the antenna a voltage will be induced in each conductor. This causes a current flow determined by the induced voltage and the impedance of the conductor.

The degree of coupling depends on the position of the conductor with respect to the antenna, assuming that all the conductors in the figure are the same length. The coupling between the antenna and conductor *IJ* is greater than in any other case, because *IJ* is close to and parallel with the antenna. Ideally, the coupling between conductor *GH* and the antenna is zero, because the voltage induced by current flowing in the left-hand side of the antenna is exactly balanced by a voltage of opposite polarity induced by the current flowing in the right-hand side. This is because the two currents are flowing in opposite directions with respect to *GH*. Complete cancellation of the induced voltages can occur, of course, only if the currents in the two halves of the antenna are symmetrically distributed with respect to the center of the antenna, and also only if every point along *GH* is equidistant from any two points along the antenna that are likewise equidistant from the center. This cannot be true of any of the other conductors shown, so a finite voltage will be induced in any conductor in the vicinity of the antenna except one perpendicular to the antenna at its center.

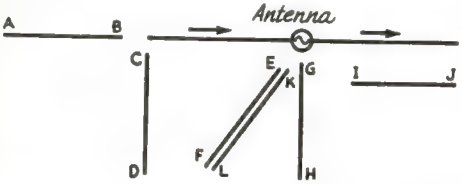


Fig. 3-47—Coupling between antenna and conductors in the antenna's field.

Transmission Line in the Antenna Field

Now consider the two conductors *EF* and *KL*, which are parallel and very close together. Except for the negligible spacing between them, the two conductors lie in the same position with respect to the antenna. Therefore, identical voltages will be induced in both, and the resulting currents will be flowing *in the same direction in both conductors.* It is only a short step to visualizing conductors *EF* and *KL* as the two conductors of a section of transmission line in the vicinity of the antenna. Because of coupling to the antenna it is not only possible but *certain* that a voltage will be induced in the two conductors of the transmission line in parallel. The resulting current flow is in the same direction in both conductors, whereas the true transmission-line currents are always flowing in opposite directions at each point along the line. These "parallel" currents are of the same nature as the current in the antenna itself, and hence are called "antenna" currents on the line. They are responsible for most of the radiation that takes place from transmission lines.

When there is an antenna current of appreciable amplitude on the line it will be found that not only are the line currents unbalanced but the apparent s.w.r. is different in each conductor, and that the loops and nodes of current in one wire do not occur at corresponding points in the other wire. Under these conditions it is impossible to measure the true s.w.r.

It should be obvious from Fig. 3-47 that only in the case of a center-fed antenna can the coupling between the line and antenna be reduced to zero. There is always some such coupling when the antenna is end fed, so there is always the possibility that antenna currents of appreciable amplitude will exist on the line, contributing further to the inherent line unbalance in the end-fed arrangement. But the center-fed system also will have appreciable antenna-to-line coupling if the line is not brought off at right angles to the antenna for a distance of at least a half wavelength.

Antenna currents will be induced on lines of any type of construction. If the line is coax, the antenna current flows only on the *outside* of the outer conductor; no current is induced *inside* the line. However, an antenna current on the outside of coax is just as effective in causing radiation as a similar current induced in the two wires of a parallel-conductor line.

Detuning the Line for Antenna Currents

The antenna current flowing on the line as a result of voltage induced from the antenna will be small if the overall circuit, considering the line simply as a single conductor, is not resonant at the operating frequency. The frequency (or frequencies) at which the system is resonant depends on the total length and whether the transmission line is grounded or not at the transmitter end.

If the line is connected to a coupling circuit that is not grounded, either directly or through a capacitance of more than a few picofarads, it is necessary to consider only the length of the antenna and line. In the end-fed arrangement, shown at A in Fig. 3-48, the line length, *L*, should not be an integral multiple or close to such a multiple of a half wave length. In the center fed system, Fig. 3-48B, the length of the line plus *one side* of the antenna should not

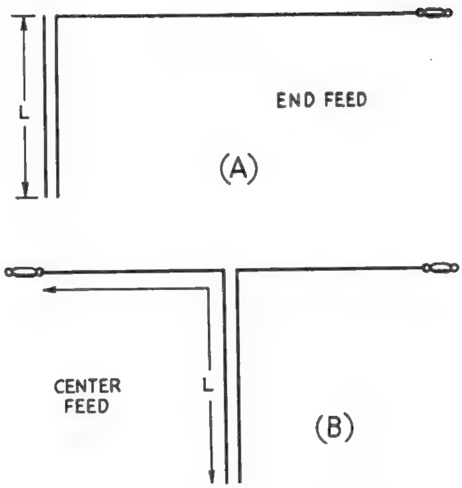


Fig. 3-48—The important length for resonance to antenna currents coupled from the antenna to the line. In the center-fed system one side of the antenna is part of the "parallel"-resonant system.

be a multiple of a half wavelength. In this case the two halves of the antenna are simply in parallel so far as resonance for the induced "antenna" current on the line is concerned, because the line conductors themselves act in parallel. When the antenna is to be used in several bands, resonances of this type should be avoided at all frequencies to be used. Fig. 3-49 shows, as solid lines along the length scale, the lengths that avoid exact resonance on frequencies from 3.5 to 29.7 Mc. These are based on the usual antenna-length formulas; the velocity factor of the line plays no part in establishing such resonances since it applies only to true transmission-line currents.

Whenever possible, it is best to choose line lengths, such as those indicated by the arrows, that fall midway in the nonresonant range. This is because the resonances are not extremely sharp. Working close to resonance, although not exactly on it, will allow an appreciable "antenna" current to flow even though it may not be as large as it would if the line were exactly resonant for it. For the same reason the line length should be chosen to fall in a range where there is a considerable distance between resonances. A length of 76 feet, for example, would be definitely less susceptible to resonance effects than a length of 96 feet.

The lengths shown in Fig. 3-49 are subject to some modification in practice. Transmission lines usually have bends, are at varying heights above ground, etc., all of which will modify the resonant frequency. It is advisable to check the system for resonance at and near all operating frequencies before assuming that the line is safely detuned for antenna currents. This can be done by temporarily connecting the ends of the line together and coupling them through a small capacitance (not more than a few $\mu\mu$f.) to a resonance indicator such as a grid-dip meter. Very short leads should be used between the meter and antenna. Fig. 3-50 shows the method. Once the resonance points are known it is a simple matter to prune the feeders to get as far away as possible from resonance at any frequency to be used.

Resonances in systems in which the coupling apparatus is grounded at the transmitter are not so easily predicted. The "ground" in such a case is usually the metal chassis of the transmitter itself, not actual ground. In the average amateur station it is not possible to get a connection to real ground without having a lead that is an appreciable fraction of wavelength long. At the higher frequencies, and particularly in the v.h.f. region, the distance from the transmitter to ground may be one wavelength or more. Probably the best plan in such cases is to make the length L in Fig. 3-48 equal to a multiple of a half-wavelength. If the transmitter has fairly large capacitance to ground, a system of this length will be effectively detuned for the fundamental and all even harmonics when grounded to the transmitter at the coupling apparatus. However, the resonance frequencies will depend on the arrangement and constants of the cou-

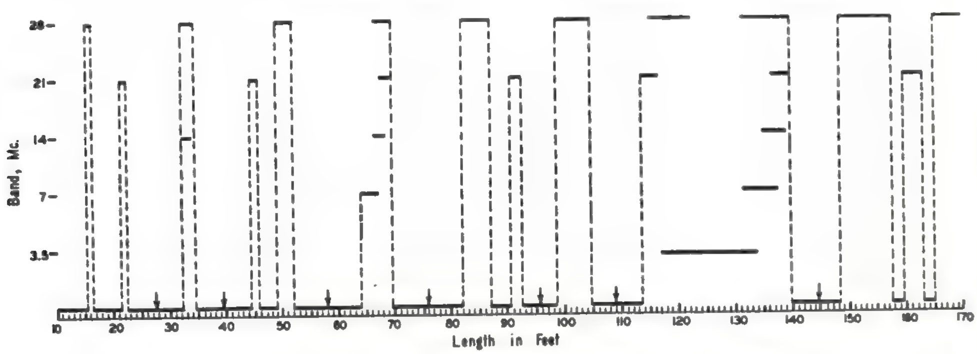


Fig. 3-49—Lengths shown by solid lines along the horizontal axis avoid exact resonance at frequencies in all amateur bands from 3.5 to 29.7 Mc., in systems where the coupling apparatus is not grounded. Best operating lengths are at the centers of the wider ranges, as shown by the arrows. These lengths correspond to L in Fig. 3-48.

pling system even in such a case, and preferably should be checked by means of the grid-dip meter. If this test shows resonance at or near the operating frequency, alternative grounds (to a heating radiator, for example) should be tried until a combination is found that detunes the whole system.

It should be quite clear, from the mechanism that produces antenna currents on a transmission line, that such currents are entirely independent of the normal operation as a true transmission line. It does not matter whether the

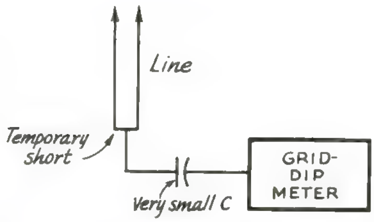


Fig. 3-50—Using a grid-dip meter to check resonance of the antenna system for antenna currents on the transmission line.

line is perfectly matched or is operated with a high standing-wave ratio. Nor does it matter what kind of line is used, air-insulated or solid-dielectric, parallel-conductor or coax. In every case, the antenna currents should be minimized by detuning the line if the line is to fulfill only its primary purpose of transferring power to the antenna.

Other Causes of Unbalance

Unbalance in center-fed systems can arise even when the line is brought away at right angles to the antenna for a considerable distance. If both halves of the antenna are not symmetrically placed with respect to near-by conductors (such as power and telephone wires, downspouting, etc.) the antenna itself becomes unbalanced and the current distribution is different in the two halves. Because of this unbalance a voltage will be induced in the line even if the line is symmetrical with respect to the antenna.

MATCHED LINES

Operating the transmission line at a low standing-wave ratio requires that the line be terminated, at its output end, in a resistive load matching the characteristic impedance of the line as closely as possible. The problem can be approached from two standpoints: selecting a transmission line having a characteristic impedance that matches the antenna resistance at the point of connection; or transforming the antenna resistance to a value that matches the Z_0 of the line selected.

The first approach is simple and direct, but its application is limited because the antenna impedance and line impedance are alike only in a few special cases. The second approach provides a good deal of freedom in that the antenna and line can be selected independently. Its disadvantage is that it is more complicated constructionally. Also, it sometimes calls for a somewhat tedious routine of measurement and adjustment before the desired match is achieved.

Operating Considerations

As pointed out earlier in this chapter, most antenna systems show a marked change in resistance when going from the fundamental to multiples of the fundamental frequency. For this reason it is usually possible to match the line impedance only on one frequency. A matched antenna system is consequently a one-band affair, in most cases. It can, however, usually be operated over a fair frequency range in a given band. The frequency range over which the standing-wave ratio is low is determined by the impedance-*vs.*-frequency characteristic of the antenna. If the change in impedance is small for a given change in frequency, the s.w.r. will be low over a fairly wide band of frequencies. However, if the impedance change is rapid (a sharply-resonant or high-Q antenna—see discussion of Q later in this chapter) the s.w.r. will also rise rapidly as the operating frequency is shifted to one side or the other of the frequency for which the antenna is resonant and for which the line is matched.

Antenna Resonance

A point that needs emphasis in connection with matching the antenna to the line is that, with the exception of a few special cases discussed later in this chapter, the impedance at the point where the line is connected must be a *pure resistance*. This means that the antenna system must be resonant at the frequency for which the line is to be matched. (Some types of long-wire antennas are exceptions, in that their input impedances are resistive over a wide band of frequencies. Such systems are essentially nonresonant.) The higher the Q of the antenna system the more essential it is that exact resonance be established before an attempt is made to match the line. This is particularly true of the close-spaced parasitic arrays described in a later chapter. With simple dipole and harmonic antennas the tuning is not so critical and it is usually sufficient to cut the antenna to the length given by the appropriate formula in Chapter Two. The frequency should be selected to be at the center of the range of frequencies (which may be the entire width of an amateur band) over which the antenna is to be used.

DIRECT MATCHING

As discussed in Chapter Two, the impedance at the center of a resonant half-wave antenna at heights of the order of ¼ wavelength and more is resistive and is in the neighborhood of 70 ohms. This is fairly well matched by transmitting-type Twin-Lead having a characteristic im-

pedance of 75 ohms. It is possible, therefore, to operate with a low s.w.r. using the arrangement shown in Fig. 3-51. No precautions are necessary beyond those already described in connection with antenna-to-line coupling.

This system is badly mismatched on *even* multiples of the fundamental frequency, since the feed in such cases is at a high-impedance point. However, it is reasonably well matched at *odd* multiples of the fundamental. For example, an antenna resonant near the low-frequency end of the 7-Mc. band will operate with a low s.w.r. over the 21-Mc. band (three times the fundamental).

The same method may be used to feed a harmonic antenna at any current loop along the wire. For lengths up to three or four wavelengths the s.w.r. should not exceed 2 to 1 if the antenna is ¼ or ½ wavelength above ground.

At the fundamental frequency the s.w.r. should not exceed about 2 to 1 within a frequency range ±2% from the frequency of exact resonnance. Such a variation corresponds approximately to the entire width of the 7-Mc. band, if the antenna is resonant at the center of the band. A wire antenna is assumed. Antennas having a greater ratio of diameter to length will have a lower change in s.w.r. with frequency.

Coaxial Cable

Instead of using Twin-Lead as just described, the center of a half-wave dipole may be fed through 75-ohm coaxial cable such as RG-11/U, as shown in Fig. 3-52. Cable having an impedance of approximately 50 ohms, such as RG-8/U, also may be used, particularly in those cases where the antenna height is such as to lower the radiation resistance of the antenna (see Chapter Two). The principle is exactly the same as with Twin-Lead, and the same remarks as to s.w.r. apply. However, there is a considerable practical difference between the two types of line. With the parallel-conductor line the system is symmetrical, but with coaxial line it is inherently unbalanced.

Stated broadly, the unbalance with coaxial line is caused by the fact that the outside of the outer conductor is not coupled to the antenna in the same way as the inner conductor and the inside of the outer conductor. The over-all result is that current will flow on the outside of the outer conductor in the simple arrangement shown in Fig. 3-53. The unbalance is rather small if the line diameter is very small compared with the length of the antenna, a condition that is met fairly well at the lower amateur frequencies. However, it is not negligible in the v.h.f. and u.h.f. range, nor should it be ignored at 28 Mc. The current that flows on the outside of the line because of this unbalance, it should be noted, does not arise from the same type of coupling as the "antenna" current previously discussed. The coupling pictured in Fig. 3-47 can still occur, *in addition*. However, the remedy is the same in both cases—the system must be detuned for currents on the outside of the line. This can be done by choosing one of the recommended lengths in Fig. 3-49 or by an actual resonance check using the method shown in Fig. 3-50.

Balancing Devices

The unbalanced coupling described in the preceding paragraph can be nullified by the use of devices that prevent the unwanted current from flowing on the outside of the coaxial line.

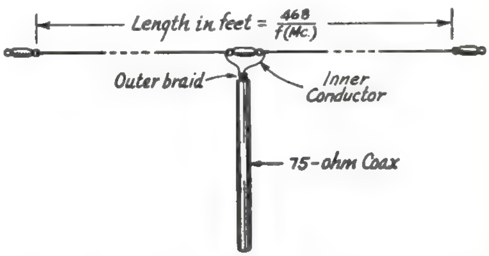


Fig. 3-52—Half-wave antenna fed with 75-ohm coaxial cable. The outside of the outer conductor of the line may be grounded for lightning protection.

This may be done either by making the current cancel itself out or by choking it off. Devices of this type fall in a class of circuits usually termed baluns, a contraction for "balanced to unbalanced."

The voltages at the antenna terminals in Fig. 3-52 are equal in amplitude with respect to ground but opposite in phase. Both these voltages act to cause a current to flow on the outside of the coax, and if the currents produced by both voltages were equal, the resultant current on the outside of the line would be zero since the currents are out of phase and would cancel each other. But since one antenna terminal is directly connected to the cable shield while the other is only weakly coupled to it, the voltage at the directly-connected terminal produces a much larger current, and so there is relatively little cancellation.

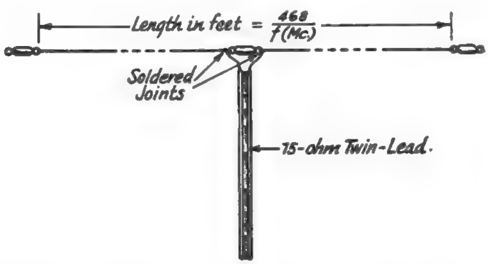


Fig. 3-51—Half-wave dipole fed with 75-ohm Twin-Lead, giving a close match between antenna and line impedance. The leads in the "Y" from the end of the line to the ends of the center insulator should be as short as possible.

The two currents could be made equal in amplitude by making a direct connection between the outside of the line and the antenna terminal that is connected to the inner conductor, but if it were done right at the antenna terminals the line and antenna would be short-circuited. However, if the connection is made through a conductor parallel to the line and a quarter wavelength long, as shown in Fig. 3-53A the second conductor and the outside of the line act as a quarter-wave "insulator" for the normal voltage and current at the antenna terminals. (This is because a quarter-wave line short-circuited at the far end exhibits a very high resistive impedance, as explained earlier in this chapter.) On the other hand, any unbalance current flowing on the outside of the line because of the direct connection between it and the antenna has a counterpart in an equal current flowing on the second conductor, because the latter is directly connected to the *other* antenna terminal. Where the two conductors are joined together at the bottom, the resultant of the two currents is zero, since they are of opposite phase. Thus no current flows on the remainder of the transmission line.

Note that the length of the extra conductor has no particular bearing on its operation in balancing out the undesired current. The length is critical only in respect to preventing the normal operation of the antenna from being upset.

Combined Balun and Matching Stub

In certain antenna systems the balun length can be considerably shorter than a quarter wavelength; the balun is, in fact, used as part of the matching system. This requires that the radiation resistance be fairly low as compared with the line Z_0 so that a match can be brought about by first shortening the antenna to make it have a capacitive reactance, and then using a shunt inductance across the antenna terminals to resonate the antenna and simultaneously raise the impedance to a value equal to the line Z_0. (See later section on matching stubs.) The balun is then made the proper length to exhibit the desired value of inductive reactance.

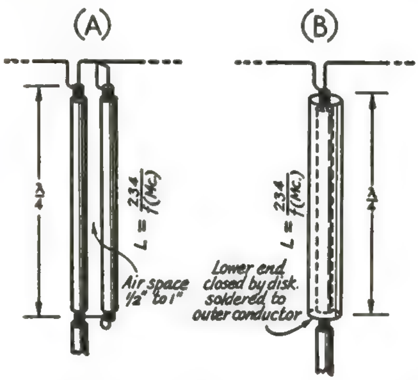


Fig. 3-53—Methods of balancing the termination when a coaxial cable is connected to a balanced antenna.

The basic matching method is shown at A in Fig. 3-54, and the balun adaptation to coaxial feed is shown at B. The matching stub in the latter case is a parallel-line section, one conductor of which is the outside of the coax between point *X* and the antenna; the other stub conductor is an equal length of wire. (A piece of coax may be used instead, as in the balun in Fig. 3-53A.) The spacing between the stub conductors can be two to three inches. The system has two advantages over Fig. 3-53: The stub is ordinarily much shorter than a quarter wavelength, and the impedance match can be adjusted by adjusting the stub length along with the antenna length. With simple coax feed, even with a quarter-wave balun, the match depends entirely on the actual antenna impedance and the Z_0 of the cable; no adjustment is possible.

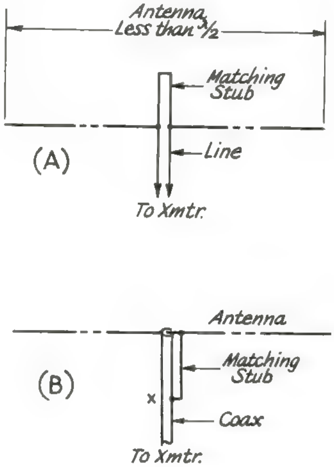


Fig. 3-54—Combined matching stub and balun. A—Basic arrangement; B—Balun arrangement achieved by using a section of the outside of the coax feed line as one conductor of a matching stub.

Adjustment

When a quarter-wave balun is used it is advisable to resonate it before connecting the antenna. This can be done without much difficulty if a grid-dip meter is available. In the system shown in Fig. 3-53A, the section formed by the two parallel pieces of line should first be made slightly longer than the length given by the formula. The shorting connection at the bottom may be installed permanently. With the grid-dip meter coupled to the shorted end, check the frequency and cut off small lengths of the shield braid (cutting both lines equally) at the open ends until the stub is resonant at the desired frequency. In each case leave just enough inner conductor remaining to make a short connection

to the antenna. After resonance has been established, solder the inner and outer conductors of the second piece of coax together and complete the connections indicated in Fig. 3-53A.

An alternative method is first to adjust the antenna length to the desired frequency, with the line and stub disconnected, then connect the balun and recheck the frequency. Its length may then be adjusted so that the over-all system is again resonant at the desired frequency.

Construction

In constructing a balun of the type shown in Fig. 3-53A the additional conductor and the line should be maintained parallel by suitable spacers. It is convenient to use a piece of coax for the second conductor; the inner conductor can simply be soldered to the outer conductor at both ends since it does not enter into the operation of the device. The two cables should be separated sufficiently so that the vinyl covering represents only a small proportion of the dielectric between them. Since the principal dielectric is air, the length of the quarter-wave section is based on a velocity factor of 0.95, approximately.

Detuning Sleeves

The detuning sleeve shown in Fig. 3-53B also is essentially an air-insulated quarter-wave line, but of the coaxial type, with the sleeve constituting the outer conductor and the outside of the coax line being the inner conductor. Because the impedance at the open end is very high, the unbalanced voltage on the coax line cannot cause much current to flow on the outside of the sleeve. Thus the sleeve acts like a choke coil in isolating the remainder of the line from the antenna. (The same viewpoint can be used in explaining the action of the quarter-wave arrangement shown at A, but is less easy to understand in the case of baluns less than ¼ wavelength long.)

A sleeve of this type may be resonated by cutting a small longitudinal slot near the bottom, just large enough to take a single-turn loop that is in turn link-coupled to the grid-dip meter. If the sleeve is a little long to start with, a bit at a time can be cut off the top until the stub is resonant.

The diameter of the coaxial detuning sleeve in B should be fairly large compared with the diameter of the cable it surrounds. A diameter of two inches or so is satisfactory with half-inch cable. The sleeve should be symmetrically placed with respect to the center of the antenna so that it will be equally coupled to both sides. Otherwise a current will be induced from the antenna to the outside of the sleeve. This is particularly important at v.h.f. and u.h.f.

In both the balancing methods shown in Fig. 3-53 the quarter-wave section should be cut to be resonant at exactly the same frequency as the antenna itself. These sections tend to have a beneficial effect on the impedance-frequency characteristic of the system, because their reactance varies in the opposite direction to that of the antenna. For instance, if the operating frequency is slightly below resonance the antenna has capacitive reactance, but the shorted quarter-wave sections or stubs have inductive reactance. Thus the reactances tend to cancel, which prevents the impedance from changing rapidly and helps maintain a low standing-wave ratio on the line over a band of frequencies.

QUARTER-WAVE TRANSFORMERS

The impedance-transforming properties of a quarter-wave transmission line can be used to good advantage in matching the antenna impedance to the characteristic impedance of the line. As described earlier, the input impedance of a quarter-wave line terminated in a resistive impedance Z_R is

$$Z_S = \frac{Z_0^2}{Z_R}$$

Rearranging this equation gives

$$Z_0 = \sqrt{Z_R Z_S}$$

This means that any value of load impedance Z_R can be transformed into any desired value of impedance Z_S at the input terminals of a quarterwave line, provided the line can be constructed to have a characteristic impedance Z_0 equal to the square root of the product of the two impedances. The factor that limits the range of impedances that can be matched by this method is the range of values for Z_0 that is physically realizable. The latter range is approximately 50 to 600 ohms.

Practically any type of line can be used for the matching section, including both air-insulated and solid-dielectric lines.

One application of this type of matching section is in matching a half-wave antenna to a 600-ohm line, as shown in Fig. 3-55. Assuming that the antenna has a resistive impedance in the vicinity of 65 to 70 ohms, the required Z_0 of the matching section is approximately 200 ohms. A section of this type can be constructed of parallel tubing, from the data in Fig. 3-25.

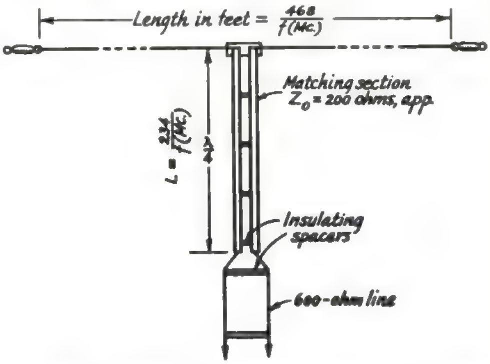


Fig. 3-55—Matching a half-wave antenna to a 600-ohm line through a quarter-wave linear transformer. This arrangement is popularly known as the "Q" matching system.

The ¼-wave transformer may be adjusted to resonance before being connected to the antenna by short-circuiting one end and coupling it inductively at that end to a grid-dip meter. The length of the short-circuiting conductor lowers the frequency slightly, but this can be compensated for by adding half the length of the shorting bar to each conductor after resonating, measuring the shorting-bar length between the centers of the conductors.

Driven Beam Elements

Another application for the quarter-wave "linear transformer" is in matching the very low antenna impedances encountered in close-spaced directional arrays to a transmission line having a characteristic impedance of 300 to 600 ohms. The observed impedances at the antenna feed point in such cases range from about 8 to 20 ohms. A matching section having a Z_0 of 75 ohms is useful with such arrays. The impedance at its input terminals will vary from approximately 700 ohms with an 8-ohm load to 280 ohms with a 20-ohm load.

Transmitting Twin-Lead is suitable for this application; such a short length is required that the loss in the matching section should not exceed about 0.6 db. even though the s.w.r. in the matching section may be almost 10 to 1 in the extreme case.

DELTA MATCHING

Among the properties of a coil-and-capacitor resonant circuit is that of transforming impedances. If a resistive impedance, Z_1 in Fig. 3-56, is connected across the outer terminals *AB* of a resonant *LC* circuit, the impedance Z_2 as viewed looking into another pair of terminals such as *BC* will also be resistive, but will have a different value depending on the mutual coupling between the parts of the coil associated with each pair of terminals. Z_2 will be less than Z_1 in the circuit shown, but this relationship will be reversed if Z_1 is connected across terminals *BC* and Z_2 is viewed from terminals *AB*.

A resonant antenna has properties similar to those of a tuned circuit. The impedance presented between any two points symmetrically placed with respect to the center of a half-wave antenna will depend on the distance between the points. The greater the separation the higher the value of impedance, up to the limiting value that exists between the open ends of the antenna. This is also suggested in Fig. 3-56. The impedance Z_A between terminals *1* and *2* is lower than the impedance Z_B between terminals *3* and *4*. Both impedances, however, are purely resistive if the antenna is resonant.

This principle is utilized in the **delta** matching system shown in Fig. 3-57. The center impedance of a half-wave dipole is too low to be matched directly by any practicable type of air-insulated parallel-conductor line. However, it is possible to find between two points a value of impedance that can be matched to such a line when a "fanned" section or delta is used to couple the line and antenna.

Available information on the delta match is based on experimental data for the case of a simple half-wave antenna coupled to a 600-ohm transmission line. The antenna length, *L*, should be based on the formula in Chapter Two, using the appropriate factor for the length/diameter ratio. The ends of the delta or "Y" should be attached at points equidistant from the center of

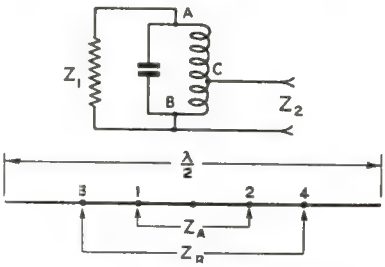


Fig. 3-56—Impedance transformation with a resonant circuit, together with antenna analogy.

the antenna, and the total distance, *A*, between them is given by

$$A \text{ (feet)} = \frac{118}{f \text{ (Mc.)}}$$

for frequencies up to and including the 28-Mc. band, and

$$A \text{ (feet)} = \frac{113}{f \text{ (Mc.)}}$$

for frequencies above 30-Mc. The length of the delta, *B*, is given by

$$B \text{ (feet)} = \frac{148}{f \text{ (Mc.)}}$$

These formulas are based on the assumption that the center impedance of the antenna is approximately 70 ohms, and will require modification if the actual impedance is very much different.

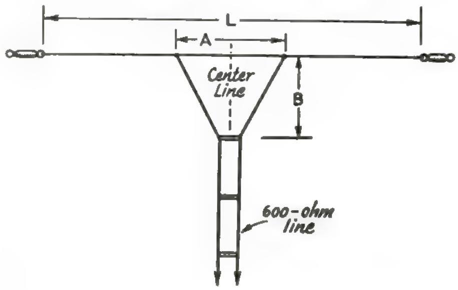


Fig. 3-57—The "delta" matching system, applied to a half-wave antenna and 600-ohm line.

The delta match can be used for matching the driven element of a directive array to a transmission line, but if the impedance of the element is low—as is frequently the case—the proper dimensions for *A* and *B* must be found by experiment.

The delta match is somewhat awkward to adjust when the proper dimensions are unknown, because both the length and width of the delta must be varied. An additional disadvantage is that there is always some radiation from the delta. This is because the conductors are not close enough together to meet the requirement (for negligible radiation) that the spacing should be very small in comparison with the wave length.

FOLDED DIPOLES

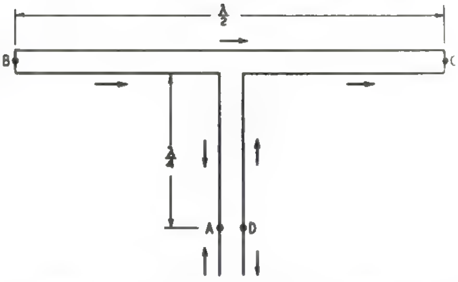


Fig. 3-58—Direction of current flow in a folded dipole and associated transmission line.

In the diagram shown in Fig. 3-58, suppose for the moment that the upper conductor between points *B* and *C* is disconnected and removed. The system is then a simple center-fed dipole, and the direction of current flow along the antenna and line at a given instant is as given by the arrows. Then if the upper conductor between *B* and *C* is restored, the current in it will flow away from *B* and toward *C*, in accordance with the rule for reversal of direction in alternate half-wave sections along a wire. However, the fact that the second wire is "folded" makes the currents in the two conductors of the antenna flow in the *same* direction. Although the antenna physically resembles a transmission line, it is not actually a line but is merely two conductors in parallel. The connections at the ends of the two are assumed to be of negligible length.

A half-wave dipole formed in this way will have the same directional properties and total radiation resistance as an ordinary dipole. However, the transmission line is connected to only *one* of the conductors. It is therefore to be expected that the antenna will "look" different, in respect to its input impedance, as viewed by the line.

The effect on the impedance at the antenna input terminals can be visualized quite readily. The center impedance of the dipole *as a whole* is the same as the impedance of a single-conductor dipole—that is, approximately 70 ohms. A given amount of power will therefore cause a definite value of current, *I*. In the ordinary half-wave dipole this current flows at the junction of the line and antenna. In the folded dipole the same total current also flows, but is equally divided between *two* conductors in parallel. The current in each conductor is therefore $I/2$. Consequently, the line "sees" a higher impedance because it is delivering the same power at only half the current. It is easy to show that the new value of impedance is equal to four times the impedance of a simple dipole. If more wires are added in parallel the current continues to divide between them and the terminal impedance is raised still more. This explanation is a simplified one based on the assumption that the conductors are close together and have the same diameter.

The two-wire system in Fig. 3-59 is an especially useful one because the input impedance is so close to 300 ohms that it can be fed directly with 300-ohm Twin-Lead or open line without any other matching arrangement, and the line will operate with a low standing-wave ratio. The antenna itself can be built like an open-wire line, that is, the two conductors can be held apart by regular feeder spreaders. TV "ladder" line is quite suitable. It is also possible to use 300-ohm line for the antenna, in addition to using it for the transmission line. Since the antenna section does not operate as a transmission line but simply as two wires in parallel, the velocity factor of Twin-Lead can be ignored in computing the antenna length.

The folded dipole has a somewhat "flatter" impedance-*vs.*-frequency characteristic than a simple dipole. That is, the reactance varies less rapidly, as the frequency is varied on either side of resonance, than with a single-wire antenna. The result is that it is possible to operate over a wider band of frequencies, while maintaining a low s.w.r. on the line, than with a simple dipole. This is partly explained by the fact that the two

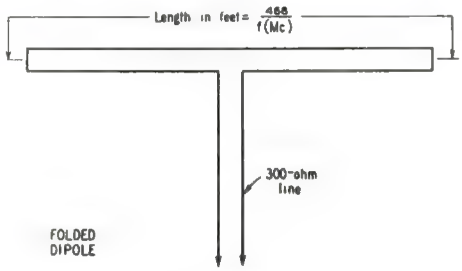


Fig. 3-59—The folded dipole.

conductors in parallel form a single conductor of greater effective diameter.

Harmonic Operation

A folded dipole will not accept power at twice the fundamental frequency, or any even multiples of the fundamental. At such multiples the folded section simply acts like a continuation of the transmission line. No other current distribution is possible if the currents in the two conductors of the actual transmission line are to flow in opposite directions. However, the antenna system as a whole can be excited at the second

harmonic if there is harmonic coupling between the transmitter tank circuit and the line, and if the overall system—considering the line conductors as being in parallel—is at or near resonance at

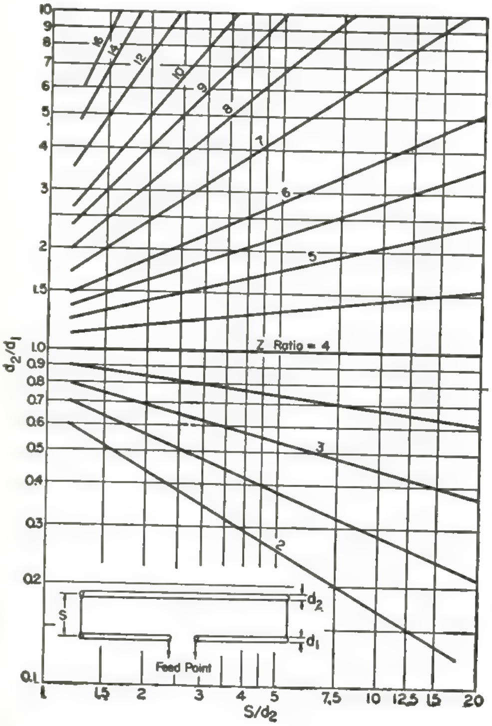


Fig. 3-60—Impedance step-up ratio for the two-conductor folded dipole, as a function of conductor diameters and spacing. Dimensions d_1, d_2 and S are shown on the inset drawing.

the second harmonic. The significant length is L in Fig. 3-48B. To avoid radiation of even harmonics, the system should be detuned by proper choice of overall length.

On the third and other odd multiples of the fundamental the current distribution is correct for operation of the system as a folded antenna. Since the radiation resistance of a 3/2-wave antenna is not greatly different from that of a half-wave antenna, a folded dipole can be operated on its third harmonic with a low s.w.r. in 300-ohm line. A 7-Mc. folded dipole consequently can be used for the 21-Mc. band as well.

Multi- and Unequal-Conductor Folded Dipoles

Larger impedance ratios than 4 to 1 are frequently desirable when the folded dipole is used as the driven element in a directive array because the radiation resistance is frequently quite low. A wide choice of impedance step-up ratios is available by varying the relative size and spacing of the conductors, and by using more than two. Fig. 3-60 gives design information of this nature for two-conductor folded dipoles and Fig. 3-61 is a similar chart for three-conductor dipoles. Fig. 3-61 assumes that the three conductors are in the same plane, and that the two that are not directly connected to the transmission line are equally spaced from the driven conductor.

In computing the length of a folded dipole using thick conductors—i.e., tubing such as is used in rotary beam antennas—it should be remembered that the resonant length may be appreciably less than that of a single-wire antenna cut for the same frequency. Besides the shortening required with thick conductors, as discussed in Chapter Two, the parallel conductors tend to act like the boundaries of a conducting sheet of the same width as the spacing between the conductors. The "effective diameter" of the folded dipole will lie somewhere between the actual conductor diameter and the maximum distance between conductors. The relatively large effective thickness of the antenna reduces the rate of change of reactance with frequency, so the tuning becomes relatively broad and the antenna length is not too critical for a given frequency. It is a good idea, however, to check the resonant frequency with a grid-dip meter in making final length adjustments. The transmission line should be disconnected and the antenna terminals temporarily short-circuited when this check is made.

Adjustment

As shown by the charts, there are two special cases where the impedance ratio is independent of the spacing between conductors. These are for a ratio of 4 with the two-conductor dipole and a

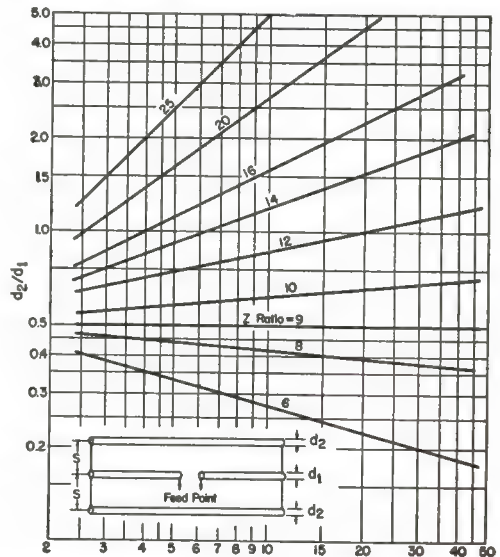


Fig. 3-61—Impedance step-up ratio for the three-conductor folded dipole. The conductors that are not directly driven must have the same diameter, but this diameter need not be the same as that of the driven conductor. Dimensions are indicated on the inset.

ratio of 9 in the three-conductor case. In all other cases the impedance ratio can be varied by adjustment of the spacing. The adjustment range is quite limited when ratios near 4 and 9, respectively, are used but increases with the departure in either direction from these "fixed" values. This offers a means for final adjustment of the match to the transmission line when the antenna resistance is known approximately but not exactly.

If a suitable match cannot be obtained by adjustment of spacing, there is no alternative but to change the ratio of conductor diameters. The impedance ratio decreases with an increase in spacing, and vice versa. Hence, if a match cannot be brought about by changing the spacing, such a change will at least indicate whether the d_2/d_1 ratio should be increased or decreased.

THE "T" AND "GAMMA"

The "T" matching system shown in Fig. 3-62 has a considerable resemblance to the folded dipole; in fact, if the distance A is extended to the full length of the antenna the system becomes an ordinary folded dipole. The "T" has considerable flexibility in impedance ratio and is more convenient, constructionally, than the folded dipole when used with the driven element of a rotatable parasitic array. Since it is a symmetrical system it is inherently balanced, and so is well suited to use with parallel-conductor transmission lines. If coaxial line is used some form of balun, as described earlier, should be installed. Alternatively, the "Gamma" form described below can be used with unbalanced lines.

The current flowing at the input terminals of the "T" consists of the normal antenna current divided between the radiator and the "T" conductors in a way that depends on their relative diameters and the spacing between them; with a superimposed transmission-line current flowing in each half of the "T" and its associated section of the antenna. Each such "T" conductor and the associated antenna conductor can be looked upon as a section of transmission line shorted at the end. Since it is shorter than ¼ wavelength it has inductive reactance; as a consequence, if the antenna itself is exactly resonant at the operating frequency the input impedance of the "T" will show inductive reactance as well as resistance. The reactance must be tuned out if a good match to the transmission line is to be secured. This can be done either by shortening the antenna to obtain a value of capacitive reactance that will reflect through the matching system to cancel the inductive reactance at the input terminals, or by inserting a capacitance of the proper value in series at the input terminals as shown in Fig. 3-64, upper drawing.

A theoretical analysis has shown that the part of the impedance step-up due to the spacing and ratio of conductor diameters is approximately the same as given for the folded dipole in Fig. 3-60. The actual impedance ratio is, however,

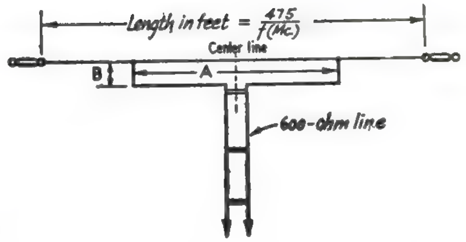


Fig. 3-62—The "T" matching system, applied to a half-wave antenna and 600-ohm line.

considerably modified by the length A of the matching section (Fig. 3-62). Means for calculation are not presently available, but the trends can be stated as follows:

1) The input impedance increases as the distance A is made larger, but not indefinitely. There is in general a distance A that will give a maximum value of input impedance, after which further increase in A will cause the impedance to decrease, the limit being when A equals the antenna length and the antenna becomes a folded dipole.

2) The distance A at which the input impedance reaches a maximum is smaller as d_2/d_1 (using the notation of Fig. 3-60) is made larger, and becomes smaller as the spacing between the conductors is increased.

3) The maximum impedance values occur in the region where A is 40 to 60 per cent of the antenna length in the average case.

4) Higher values of input impedance can be realized when the antenna is shortened to cancel the inductive reactance of the matching section.

Simple Dipole Matching

For a dipole having an approximate impedance of 70 ohms, the matching-section dimensions for matching a 600-ohm line are given by the following formulas:

$$A \text{ (feet)} = \frac{180.5}{f \text{ (Mc.)}}$$

$$B \text{ (inches)} = \frac{114}{f \text{ (Mc.)}}$$

These formulas apply for wire antennas with the matching section made of the same size wire. With an antenna element of different impedance, or for matching a line having a Z_0 other than 600 ohms, the matching-section dimensions can be determined experimentally.

The "Gamma"

The "Gamma" arrangement shown in Fig. 3-63 is an unbalanced version of the "T" suitable for use with coaxial lines. Except for the fact that the matching section is connected between the center and one side of the antenna, the remarks above about the behavior of the "T" apply equally well. The inherent reactance of the matching section can be cancelled either by shortening the antenna appropriately or by

using the resonant length and installing a capacitor *C* as shown in the lower drawing of Fig. 3-64.

Adjustment

Because of lack of an adequate method for taking into account the effect of the length of the matching section on the input impedance, the proper constants for the "T" and "Gamma" must be determined experimentally.

The use of the series capacitor as shown in Fig. 3-64 is recommended for ease of adjustment. With a trial position of the tap or taps on the antenna, measure the s.w.r. on the transmission line and adjust *C* (both capacitors simultaneously in the case of the "T") for minimum s.w.r. If it is not close to 1 to 1, try another tap position and repeat. It may be necessary to try another size of conductor for the matching section if satisfactory results cannot be secured. Changing the spacing will show which direction to go in this respect, just as in the case of the folded dipole discussed in the preceding section.

MATCHING STUBS

As explained earlier in this chapter, a terminated transmission line less than ¼ wavelength long has an input impedance that is both resistive and reactive. The equivalent circuit of the line input impedance can be formed either of

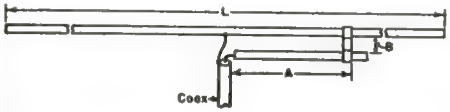


Fig. 3-63—The "Gamma match," as used with tubing elements. The transmission line may be either 52-ohm or 75-ohm coax.

resistance and reactance in series or resistance and reactance in parallel. Fig. 3-65A shows such a line terminated in a resistance less than its characteristic impedance, together with the equivalent circuits of the input impedance.

Depending on the line length, the resistive component, R_S; can have any value between Z_R (when the line has zero length) and Z_0^2/Z_R (when the line is exactly ¼ wave long). The same thing is true of R'_S. (R_S and R'_S do not however, have the same values at the same line length, other than zero and ¼ wavelength.) With either equivalent there is some line length that will give a value of R_S or R'_S equal to the characteristic impedance of the line. However, there always will be reactance along with the resistance. But if provision is made for canceling or "tuning out" this reactive part of the input impedance only the resistance will remain. Since this resistance is equal to a Z_0 a transmission line of the same characteristic impedance connected to terminals *AB* will be properly matched.

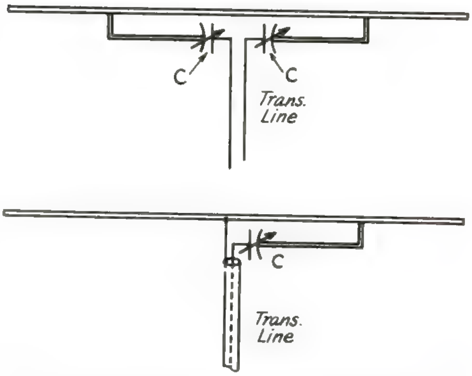


Fig. 3-64—Series capacitors for tuning out residual reactance with the "T" and "Gamma" matching systems. A maximum capacitance of 150 μμf. in each condenser should provide sufficient adjustment range, in the average case, for 14-Mc. operation. Proportionately smaller capacitance values can be used on higher-frequency bands. Receiving-type plate spacing will be satisfactory for power levels up to a few hundred watts.

Tuning out the reactance in the equivalent series circuit requires that a reactance of the same value as X_S, but of opposite kind, be inserted in series with the line. This is shown at B in Fig. 3-65. Tuning out the reactance in the equivalent parallel circuit requires that a reactance of the same value as X'_S but of the opposite kind be connected across terminals *AB*. This is shown at Fig. 3-65C.

Corresponding circuits and relationships for the case where the line is terminated in a resistive load greater than its characteristic impedance are shown in Fig. 3-66. Aside from the fact that the reactances are of the opposite kind to those in Fig. 3-65, all of the foregoing remarks apply equally well.

In practice it is convenient to use the parallel equivalent circuit. The transmission line is simply connected to the load (which of course is usually a resonant antenna) and then a react-

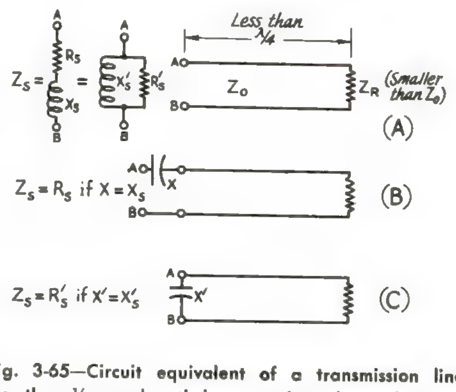


Fig. 3-65—Circuit equivalent of a transmission line less than ¼ wavelength long terminated in a load resistance lower than Z_0, and methods of compensating for the reactive component of the input impedance.

ance of the proper value is connected across the line at the proper distance from the load. From this point back to the transmitter there are no standing waves on the line. A convenient type of reactance to use is a section of transmission line less than one-quarter wavelength long, either open-circuited or short-circuited depending on whether capacitive reactance or inductive reactance is called for. Reactances formed from sections of transmission line are called **matching stubs**, and are designated as **open** or **closed** depending on whether the free end is open- or short-circuited. The two types of matching stubs are shown in the sketches of Fig. 3-67.

The distance from the load to the stub (dimension *A* in Fig. 3-67) and the length of the stub, *B*, depend on the characteristic impedances of the line and stub and on the ratio of Z_R to Z_0. Since the ratio of Z_R to Z_0 is also the standing-wave ratio in the absence of matching, the dimensions are a function of the standing-wave ratio. If the line and stub have the same Z_0, dimensions *A* and *B* are dependent on the standing-wave ratio only. Consequently, if the standing-wave ratio can be measured before the stub is installed, the stub can be properly located and its length determined even though the actual value of load impedance is not known.

Typical applications of matching stubs are shown in Fig. 3-68. From inspection of these

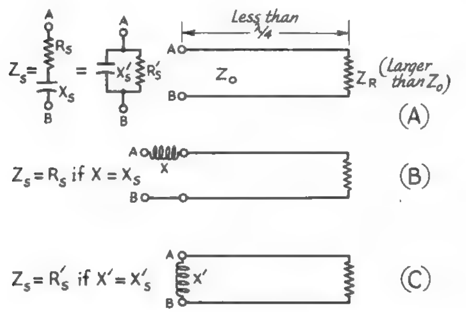


Fig. 3-66—Circuit equivalent of a transmission line less than ¼ wavelength long terminated in a load resistance greater than Z_0, and methods of compensating for the reactive component of the input impedance.

drawings it will be recognized that when an antenna is fed at a current loop, as in Fig. 3-68A, Z_R is less than Z_0 (in the average case) and therefore an open stub is called for, installed within the first quarter wavelength of line measured from the antenna. Voltage feed, as at B and C, corresponds to Z_R greater than Z_0 and therefore requires a closed stub.

Curves showing the length of the stub and its distance from the load, measured in terms of wavelength, are given in Figs. 3-69 and 3-70. These curves are based on the assumption that the load is a pure resistance, and that the characteristic impedances of the line and stub are identical. The equations from which the curves are constructed are

$$\tan A = \sqrt{S.W.R.}$$

$$\cot B = \frac{S.W.R. - 1}{\sqrt{S.W.R}}$$

for the closed stub when Z_R is greater than Z_0 and

$$\cot A = \sqrt{S.W.R.}$$

$$\tan B = \frac{S.W.R. - 1}{\sqrt{S.W.R.}}$$

for the open stub when Z_R is less than Z_0. In these equations the lengths *A* and *B* must be

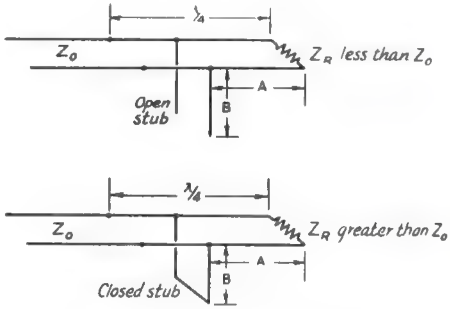


Fig. 3-67—Use of open or closed stubs for canceling the parallel reactive component of input impedance.

expressed in electrical degrees. The formula for converting length in wavelengths to electrical degrees is

Length (degrees) =
360 × Length in wavelengths

In using the curves or the above equations it must be remembered that the wavelength along the line is not the same as in free space. If an open-wire line is used the velocity factor of 0.975 will apply. When solid-dielectric line such as Twin-Lead is used the free-space wavelength as given by the curves must be multiplied by the appropriate velocity factor to obtain the actual length of *A* and *B*. The formula for the actual length is

$$A \text{ (feet)} = \frac{985 \times V}{f \text{ (Mc.)}} \times A \text{ (wavelengths)}$$

$$B \text{ (feet)} = \frac{985 \times V}{f \text{ (Mc.)}} \times B \text{ (wavelengths)}$$

where *V* is the velocity factor of the line of which sections *A* and *B* are constructed.

Although the curves and equations do not apply when the characteristic impedances of the line and stub are not the same, this does not mean that the line cannot be matched under such conditions. The stub can have any desired characteristic impedance if its length is chosen so that it has proper value of reactance. Once the length is determined for a particular case from Figs. 3-69 and 3-70 the corresponding value of reactance can be found from Fig. 3-23

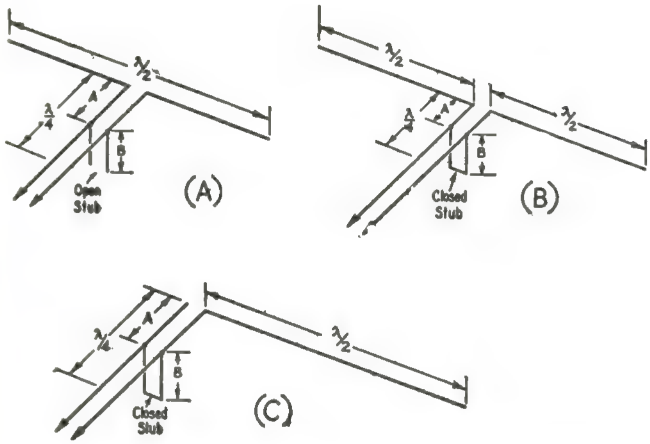


Fig. 3-68—Application of matching stubs to common types of antennas.

and a new length found from the same figure for a stub of the desired Z_0.

For example, suppose that it is determined from Fig. 3-69 that the length of an open stub is 0.16 wavelength when the stub and line have the same characteristic impedance. Assume that this Z_0 is 600 ohms. From Fig. 3-23, X/Z_0 for an open-circuited transmission line 0.16 wavelength long (57.6 degrees) is 0.65, so $X = 0.65 \times 600 = 390$ ohms. If we want to make the stub from 300-ohm Twin-Lead line, $X/Z_0 = 390/300 = 1.3$. The value 1.3 on the open-circuited line curve corresponds to a length of 37.5 degrees or 0.104 wavelength. The *actual* length of the 300-ohm Twin-Lead stub would be $0.104 \times V = 0.104 \times 0.82 = 0.085$ wavelength.

In using matching stubs it should be noted that the length and location of the stub should be based on the standing-wave ratio *at the load.* If the line is long and has fairly high losses, measuring the s.w.r. at the input end will not give the true value at the load. This point was discussed earlier in this chapter in the section on attenuation.

Reactive Loads

In this discussion of matching stubs it has been assumed that the load is a pure resistance. This is the most desirable condition, since the antenna that represents the load preferably should be tuned to resonance before any attempt is made to match the line. Nevertheless, matching stubs can be used even when the load is considerably reactive. A reactive load simply means that the loops and nodes of the standing waves of voltage and current along the line do not occur at integral multiples of ¼ wavelength from the load. To use the curves of Figs. 3-69 and 3-70 it is necessary to find a point along the line at which a current loop or node occurs. Then Fig. 3-69 gives the stub length and distance *toward the transmitter* from a current loop. Fig. 3-70 gives the stub length and distance *toward the transmitter* from a current node.

Stubs on Coaxial Lines

The principles outlined in the preceding section apply also to coaxial lines. The coaxial cases corresponding to the open-wire cases shown in

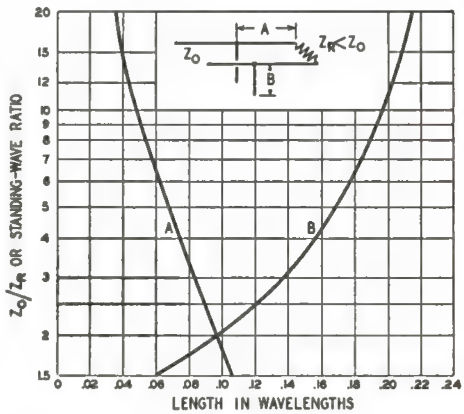


Fig. 3-69—Position and length of open stub as a function of Z_0/Z_R or standing-wave ratio.

Fig. 3-67 are given in Fig. 3-71. The curves of Figs. 3-69 and 3-70 may be used to determine the dimensions *A* and *B*. In a practical installation the junction of the transmission line and stub would be a Tee connector.

A special case of the use of a coaxial matching stub in which the stub is associated with the transmission line in such a way as to form a balun has been described earlier in this chapter (Fig. 3-54). The principles used are those just described. The antenna is shortened to introduce just enough reactance at its input terminals to permit the matching stub to be connected at

that point, rather than at some other point along the transmission line as in the general cases discussed here. To use this method the antenna resistance must be lower than the Z_0 of the main transmission line, since the resistance is transformed to a higher value. In beam antennas this will nearly always be the case.

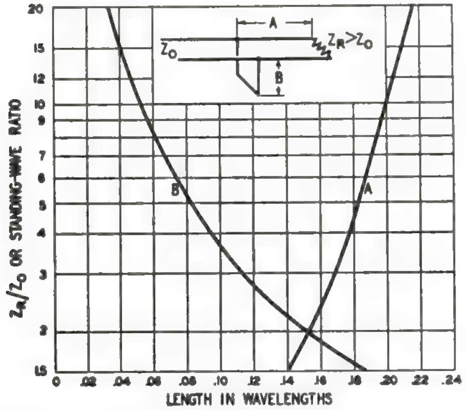


Fig. 3-70—Position and length of closed stub as a function of Z_R/Z_0 or standing-wave ratio.

Matching Sections

If the three antenna systems in Fig. 3-68 are redrawn in somewhat different fashion, as shown in Fig. 3-72, there results a system that differs in no consequential way from the matching stubs previously described, but in which the stub formed by *A* and *B* together is called a "quarter-wave matching section." The justification for this is that a quarterwave section of line is similar to a resonant circuit, as described earlier in this chapter, and it is therefore possible to use it to transform impedances by tapping at the appropriate point along the line.

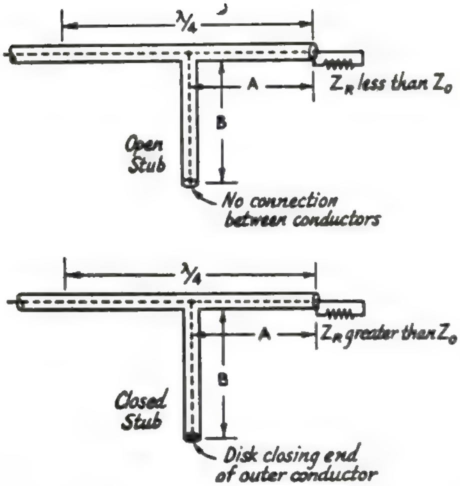


Fig. 3-71—Open and closed stubs on coaxial lines.

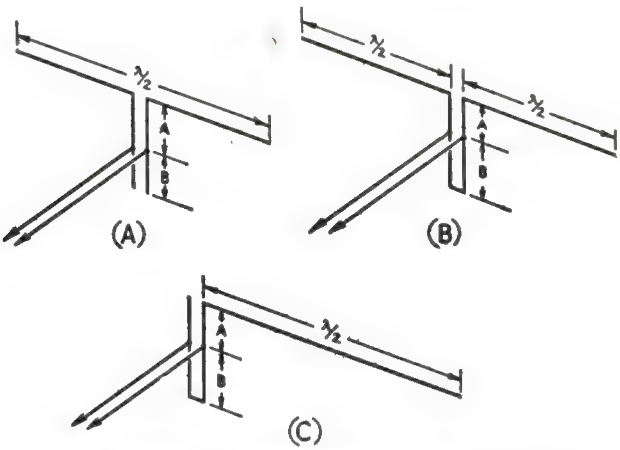


Fig. 3-72—Application of matching sections to common antenna types.

Figs. 3-69 and 3-70 give design data for matching sections, A being the distance from the antenna to the point at which the line is connected, and $A + B$ being the total length of the matching section. The curves apply only in the case where the characteristic impedances of the matching section and transmission line are the same. Equations are available for the case where the matching section has a different Z_0 than the line, but are somewhat complicated and will not be given here, since it is generally possible to make the line and matching section similar in construction.

Adjustment

In the experimental adjustment of any type of matched line it is necessary to measure the standing-wave ratio with fair accuracy in order to tell when the adjustments are being made in the proper direction. In the case of matching stubs, experience has shown that experimental adjustment is unnecessary, from a practical standpoint, if the s.w.r. is first measured with the stub not connected to the transmission line, and the stub is then installed according to the design data of Figs. 3-69 and 3.70.

MATCHING WITH LUMPED CONSTANTS

It was pointed out earlier that the purpose of a matching stub is to cancel the reactive component of line impedance at the point of connection. In other words, the stub is simply a reactance of the proper kind and value shunted across the line. It does not matter what physical shape this reactance takes. It can be a section of transmission line or a "lumped" inductance or capacitance as desired. Fig. 3-73A shows the case of a line terminated in a load impedance less than the line characteristic impedance of the line, calling for an open (capacitive) stub for impedance matching. A capacitor having the same value of reactance can be used just as well, as shown in Fig. 3-73B. There are cases where, from an installation standpoint, it may be considerably more convenient to use a capacitor in place of a stub. If a variable capacitor is used for X_C it becomes possible to adjust the capacitance to the exact value required.

The proper value of capacitance may be determined with the aid of Figs. 3-69 and 3-23. As an example, suppose that the antenna is a close-spaced array fed by a 300-ohm line, and that the standing-wave ratio at the load has been determined to be 15 to 1. From 3-69 dimension A is 0.04 wavelength and B is 0.206 wavelength. From Fig. 3-23, 0.206 wavelength corresponds to a value of X_C/Z_0 of 0.28, so $X_C = 0.28 \times 300 = 84$ ohms. If the frequency is 14.2 Mc., for instance, 84 ohms correspond to a capacitance of 134 pf. A 150-pf. variable capacitor connected across the line 0.04 wavelength from the antenna terminals would provide ample adjustment range. The r.m.s. voltage across the capacitor is $E = \sqrt{PZ_0}$ and for 500 watts, for example, would be $E = \sqrt{500 \times 300} = 386$ volts. The peak voltage is 1.41 times the r.m.s. value, or 545 volts. With 100 per cent amplitude modulation the peak would be twice as large.

Impedance-Matching Networks

The impedance-transforming properties of a coil-and-capacitor resonant circuit also can be used for matching the load to the transmission line. If the line is disconnected from terminals *1-2* in Fig. 3-73C, we have an ordinary *LC* circuit with resistance in series with the inductance. By proper choice of values for X_L and X_C, the impedance between terminals *1-2* can be made purely resistive and of a value equal to the characteristic impedance of the transmission line, provided the load resistance, Z_R, is less than the characteristic impedance of the line. The reactances required to meet this condition are given by

$$X_L = Z_R \sqrt{\frac{Z_0}{Z_R} - 1} \text{ ohms}$$

and

$$X_C = \frac{Z_0}{\sqrt{\frac{Z_0}{Z_R} - 1}} \text{ ohms.}$$

X_L and X_C do not have the same value unless the Q of the matching circuit is high. This is not generally the case because an extremely large ratio of Z_0 *to* Z_R (that is, the s.w.r. prior to matching) is required for high Q. The s.w.r. has to be practically 100 to 1 for a Q of 10, for example.

As shown in Fig. 3-73C, the inductive reactance should be divided into two equal parts, one in each side of the circuit, when a balanced line is used. Each coil, in other words, should have

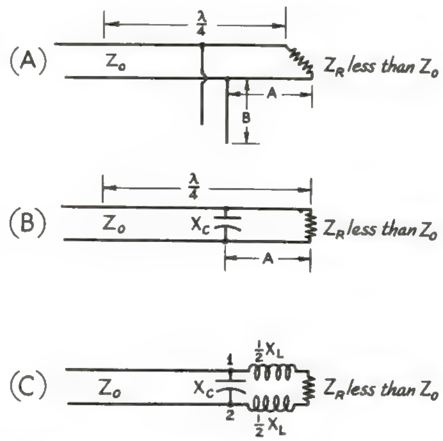


Fig. 3-73—The open stub (A) may be replaced by a capacitor of the same reactance (B) connected at the same point on the line. A matching circuit using lumped constants is shown at C.

half the total inductance, provided the two are mounted so there is no mutual inductance. The reactance values can be converted to inductance and capacitance by the usual formulas or charts.

Both systems described above can be used for matching the line to a load impedance higher than the characteristic impedance. Thus an inductor can be shunted across the line instead of the capacitor in Fig. 3-73B, at the point where a closed stub would be required. The same procedure is used for finding the required reactance, but Fig. 3-70 is used instead of Fig. 3-69, and the X_L/Z_0 curve is used in Fig. 3-23. In the circuit of Fig. 3-73C the inductive and capacitive reactances are interchanged, with one-half the total capacitive reactance (twice the required capacitance) in series with each side of the line and with the inductive reactance across the line. The formulas given above are likewise interchanged. In practice, however, this system is most likely to be used when the load impedance is less than the characteristic impedance of the line.

INDUCTIVE COUPLING

It is possible to match impedances by using inductively-coupled resonant circuits connected to the line and antenna. Although such a method has little application in most antenna systems, it it quite useful when a line is to be matched to a rotatable beam antenna. In such a case it avoids the necessity for any direct metallic connection between the antenna and line, and so allows continuous rotation without slip rings or other forms of contactor.

In its usual form the coupling is between two metallic loops or rings of rather large diameter constructed of copper tubing. These constitute the inductances in the primary (line) and secondary (antenna) circuits. The reactance of the secondary ring is tuned out by inserting capacitors C_2 and C_3 in Fig. 3-74, having a total reactance equal to that of the secondary loop. This permits the antenna or driven element to be the proper length for resonance at the operating frequency.

To obtain sufficient coupling between the two circuits it is necessary that their Qs be fairly high. The impedance of a center-fed parasitic-beam antenna element is always so low that the secondary circuit must be series resonant to operate at all, and the lower the antenna impedance the better, from the standpoint of attainable Q. On the other hand, the design of the primary circuit depends on the characteristic impedance of the transmiision line and the reactance that it is possible to obtain in L_1. A single-turn loop approximately one foot in diameter will have a reactance in the neighborhood of 75 ohms at 14 Mc. In a parallel-resonant circuit terminating a 600-ohm line this will give a Q of roughly $600/75 = 8$, when resonated by a capacitor of the same reactance. This is a satisfactory value for adequate coupling, particularly if the same size of loop is used in the secondary circuit with an antenna having an impedance of 20 ohms or less.

To use a series-resonant circuit in the primary side it would be necessary that the transmission line be 50- or 75-ohm coaxial cable or Twin-

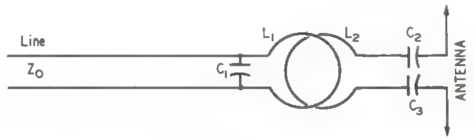


Fig. 3-74—Matching with inductively-coupled resonant circuits. This method is useful with low-impedance rotatable directive arrays.

Lead, and that the reactance of the primary loop be in the neighborhood of 500 ohms. This would require a coil of several turns, which is somewhat undesirable from a constructional standpoint. Consequently, the system is chiefly used with an open-wire transmission line.

C_1, C_2 and C_3 should be adjustable. As a preliminary step, L_1C_1 should be tuned to resonance at the operating frequency with the line disconnected and L_2 removed. (Resonance may be determined with a calibrated grid-dip meter or other suitable equipment.) $L_2C_2C_3$ also should be tuned to resonance, with the antenna disconnected and replaced by a short-circuit, and with L_1 removed. C_2 and C_3 should be adjusted to have the same capacitance at resonance. The antenna should likewise be resonated with $L_2C_2C_3$ removed. The complete system may then be connected up as shown in Fig. 3-74, leaving the capacitor settings untouched. Provision should be made for varying the coupling between L_1 and L_2. Starting with loose coupling, the transmitter is then connected to the transmission line and the coupling between L_1 and L_2 increased until the s.w.r. is as low as possible. If it is not close to unity, small adjustments may be made to the variable capacitor to make it as low as possible.

In a rotating system using this method, every effort should be made to keep the spacing between L_1 and L_2 constant as the antenna is turned. If there is any variation in spacing the power input to the antenna will vary with the position of the antenna.

It is possible to dispense with capacitors C_2 and C_3 and simply connect L_2 directly to the antenna terminals. This detunes the driven element, but the secondary circuit as a whole can be brought back to resonance by shortening the antenna until it shows a capacitive reactance equal to the inductive reactance of L_2. Alternatively, the antenna may be left at the normal resonant length, in which case reactance will be coupled into the primary circuit. This may be tuned out, within limits, by appropriate adjustment of C_1. In general, this method will require closer coupling between L_1 and L_2; also, it is likely to be more sensitive to changes in coupling as the antenna is rotated, and the

proper circuit constants are difficult to determine.

With loops as described above, approximately 150 pf. is required at C_1 and 300 pf. each at C_2 and C_3, for 14-Mc. operation. Loops of the same size at 28 Mc. will require one fourth these values of capacitance.

GROUND-PLANE ANTENNAS

Although the discussion of matching methods in this section has so far been based on symmetrical antennas, as represented by the half-wave dipole, the same principles apply to an unsymmetrical system such as the grounded antenna or the ground-plane antenna. In the case of the quarter-wave ground-plane antenna a straightforward design procedure for matching is possible because the radiation resistance is essentially independent of the physical height of the system (provided the radiator is reasonably clear of other conductors in the vicinity) and there is no ground-connection resistance to be included in the total resistance to be matched.

The ground-plane antenna lends itself well to direct connection to coaxial line, so this type of line is nearly always used. Several matching methods are available. If the antenna length can be adjusted to resonance, the stub matching system previously described is convenient. This is shown schematically in Fig. 3-75. The length of the radiator, L_a, is determined by applying

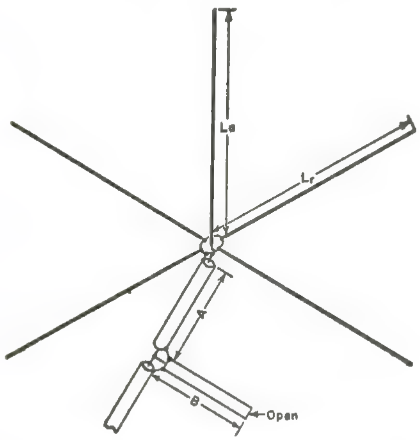


Fig. 3-75—Stub matching for the ground-plane antenna.

the length correction factor, K_a, given in Fig. 3-76, to the free-space quarter wavelength represented by the median operating frequency. To determine M, the ratio of a free-space half wavelength to the conductor diameter, the following formula may be used:

$$M = \frac{5906,}{FD}$$

where F = frequency in megacycles,
D = conductor diameter in inches.

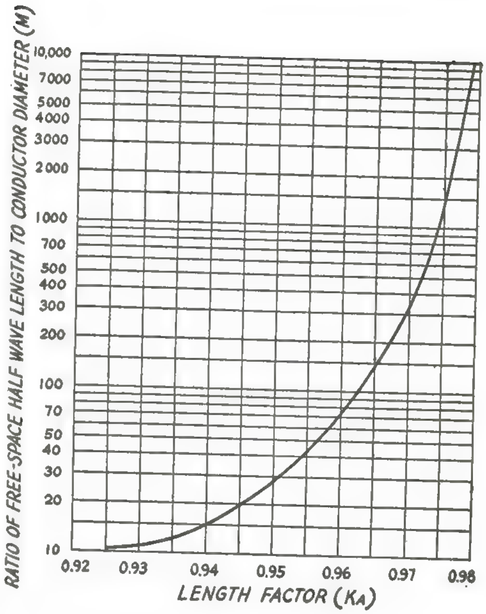


Fig. 3-76—Length factor for the range of conductor diameters used in practice. This curve applies to either quarter-wave (grounded or ground-plane antennas) or half-wave antennas. For the quarter-wave antenna the physical length is equal to the factor as given by the chart, for the particular value of M used, multiplied by a free-space quarter wavelength.

The proper length, L_r, for the radials is determined in the same way, but using a factor K_a appropriate to the diameter of the conductors used for the radials (the radiator and radials can be of different diameters). This length should be measured from the center line of the radiator to the tip of the radial.

Stub Matching

With the system properly resonant, the load to be matched to the transmission line is the radiation resistance as given by Fig. 3-77. The value of resistance determined from this curve gives the standing-wave ratio:

$$S.W.R. = \frac{Z_0}{R_r}$$

where Z_0 is the characteristic impedance of the coaxial line to be used and R_r is from Fig. 3-77. Alternatively, the actual s.w.r. can be measured with a bridge. The length of the open matching stub and distance from the antenna can then be taken from Fig. 3-69. These are electrical, not physical lengths, and must be computed on the basis of the velocity factor of the transmission line (0.66 for solid-dielectric coaxial lines).

A "tee" connector can be used for coupling the stub to the main transmission line. If the stub is exposed to the weather, the open end should be carefully sealed by rubber tape or similar means to prevent moisture from entering.

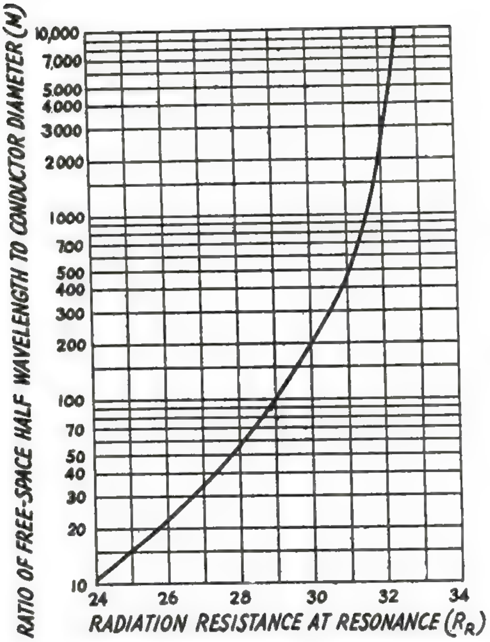


Fig. 3-77—Radiation resistance of a quarter-wave (grounded or ground-plane) antenna as a function of M. The values given apply when the antenna is adjusted to the resonant length as given by the correction factor in Fig. 3-76. Resistance values should be multiplied by 2 for center-fed half-wave antennas.

A second method of matching, particularly convenient for small antennas (28 Mc. and higher frequencies) mounted on top of a supporting mast or pole, requires shortening the antenna to the high-frequency side of resonance so that it shows a particular value of capacitive reactance at its base. The antenna terminals are then shunted by an inductive reactance, which may have the physical form either of a coil or a closed stub, to restore resonance and simultaneously transform the radiation resistance to the proper value for matching the transmission line. This method is shown in Fig. 3-78.

The first step in the procedure is to compute M and determine the resonant lengths of the radiator (L_a) and radials (L_r), and also the radiation resistance (R_r) at the resonant length, as previously described. Then, using the length factor, K_a, for the radiator, find the reactance change per 1 per cent change in length (K_x) from Fig. 3-79. Since the antenna is to be shortened, the various values must be modified appropriately. The actual radiation resistance, after the antenna is properly shortened, will be

$$R_0 = R_r - \frac{Z_0}{4R_r} \text{ ohms,}$$

where R_0 = radiation resistance after shortening, Z_0 = characteristic impedance of transmission line to be matched.

The proper value of capacitive reactance in the shortened antenna is given by

$$X_a = SR_0 \text{ ohms,}$$

where X_a = capacitive reactance of antenna, and

$$S = \sqrt{\frac{Z_0}{R_0} - 1}.$$

The antenna length that gives the proper capacitive reactance is

$$L_a = \frac{2953 K_a K_b}{F} \text{ inches,}$$

where L_a = required antenna length, and

$$K_b = 1 - \frac{X_a}{100 K_x}.$$

The inductive reactance required to be shunted across the antenna terminals for matching the transmission line is given by

$$X_s = \frac{Z_0}{S} \text{ ohms,}$$

If desired, this reactance may take the form of a coil having the requisite inductance, which is given by the formula

$$L = \frac{X_s}{6.28F}$$

where L is the inductance in microhenrys and F

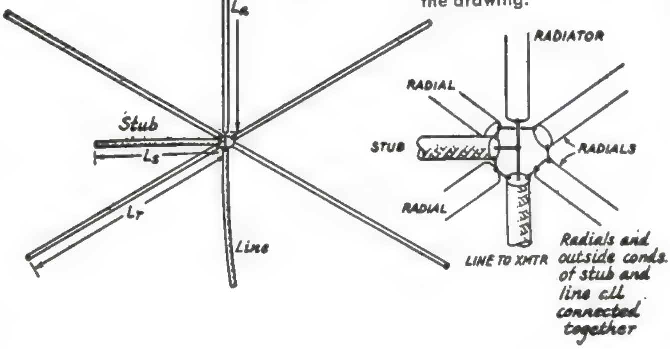


Fig. 3-78—Matching system for ground-plane antenna using inductive shunt across antenna terminals. At the higher frequencies the required inductance is conveniently supplied by a short-circuited stub as shown in the drawing.

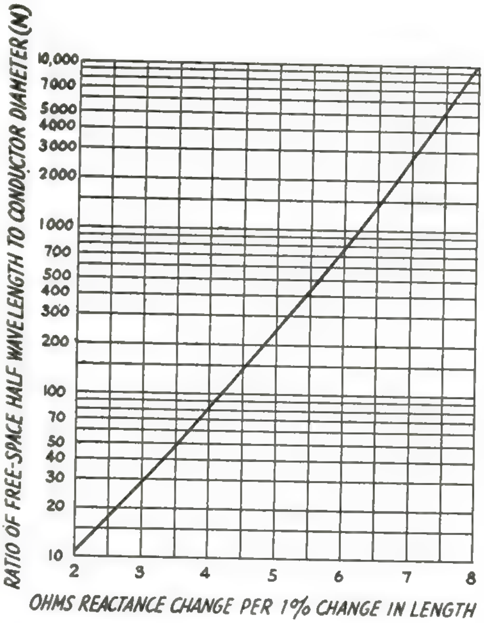


Fig. 3-79—Reactance change per unit change in antenna length, as a function of M. Values given are for quarter-wave (grounded or ground-plane) antennas and should be multiplied by 2 for center-fed half-wave antennas. Reactance is inductive if the antenna is longer than the resonant length; capacitive if shorter.

is the frequency in megacycles. In case the inductive reactance is to be supplied by a shorted stub as shown in Fig. 3-78, the stub length is

$$L_s = \frac{32.8VL}{F} \text{ inches,}$$

where L_s = stub length,
V = velocity factor of line used in stub,
L = length of stub in electrical degrees having required X_s.

L is equal to the angle whose tangent is X_sZ_s, where Z_s is the characteristic impedance of the stub.

Tapped-Coil Matching

The matching arrangement shown in Fig. 3-80 is a more general form of the method just described, in that it does not require adjusting the radiator height to an exact value. The radiator must be shortened so that the system will show capacitive reactance, but any convenient amount of shortening can be used. This system is particularly useful on lower frequencies where it may not be possible to obtain a height approximating a quarter wavelength.

The antenna impedance is matched to the characteristic impedance of the line by adjusting the taps on L. As a preliminary adjustment, before attaching the line tap (*2*), the radiator tap (*1*) may be set for resonance at the operating frequency as indicated by a grid-dip meter coupled to L. The line tap (*2*) is then moved along the coil to find the point that gives minimum standing-wave ratio as indicated by an s.w.r. bridge. To bring the s.w.r. down to 1 to 1 it will usually then be necessary to make a small readjustment of the radiator tap (*1*) and perhaps further "touch up" the line tap (*2*), since the adjustments interlock to some extent.

This method is equivalent to tapping down on a parallel-resonant circuit to match a low value of resistance to a higher value connected across the whole circuit. The antenna impedance can be represented by a capacitance in parallel with a resistance which is much higher than the radiation resistance as given by Fig. 3-77. The transformation of resistance comes about by utilizing the parallel equivalent of the radiation resistance and capacitive reactance in series, using the relationships given in Fig. 3-16.

Matching by Length Adjustment

Still another method of matching may be used when the antenna length is not fixed by other considerations. As shown in Chapter Two, the radiation resistance as measured at the base of a ground-plane antenna increases with the antenna height, and it is possible to choose a height such that the base radiation resistance will equal the Z_0 of the transmission line to be used. The heights of most interest are a little over 100 degrees (0.28 wavelength), where the resistance

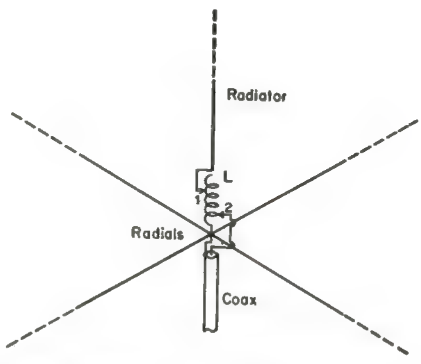


Fig. 3-80—Matching to ground-plane antenna by tapped coil. This requires that the antenna (but not radicals) be shorter than the resonant length.

is approximately 52 ohms, and about 113 degrees (0.32 wavelength), where the resistance is 75 ohms, to match the two common types of coaxial line. These heights are quite practicable for ground-plane antennas for 14 Mc. and higher frequencies. The lengths (heights) in degrees as given above do not require any correction for length/diameter ratio; i.e., they are free-space lengths.

Since the antenna is not resonant at these lengths, its input impedance will be reactive as well as resistive. The reactance must be tuned out in order to make the line see a purely resistive load equal to its characteristic impedance.

This can be done with a series capacitor of the proper value, as indicated in Fig. 3-81. The approximate value of capacitive reactance re-

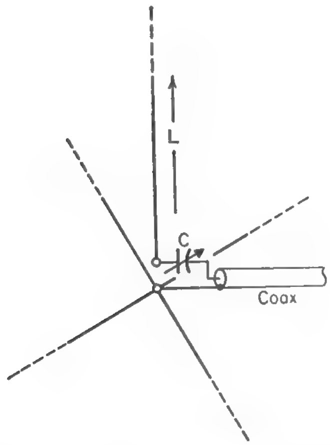


Fig. 3-81—Matching to ground-plane antenna by choice of length to give a base radiation resistance equal to the line Z_0 and tuning out reactance. The antenna height L and capacitor C are discussed in the text.

quired, for antennas of typical length/diameter ratio, is about 100 ohms for the 52-ohm case and about 200 ohms for the 75-ohm case. The corresponding capacitance values for the frequency in question can be found with the aid of the chart shown earlier in this chapter. Variable capacitors of sufficient range should be used.

The adjustment of this system requires only that the capacitance be varied until the lowest possible s.w.r. is obtained. If the lengths mentioned above are used, the s.w.r. should be close enough to 1 to 1 to make a fine adjustment of the length unnecessary.

BANDWIDTH

Although more properly a subject for discussion in connection with antenna fundamentals, the band width of the antenna is considered here because as a practical matter the change in antenna impedance with frequency is reflected as a change in the standing-wave ratio on the transmission line. Thus when an antenna is matched to the line at one frequency—usually in the center of the band of frequencies over which the antenna is to be used—a shift in the operating frequency will be accompanied by a change in the s.w.r. This would not occur if the antenna impedance were purely resistive and constant regardless of frequency, but unfortunately no practical antennas are that "flat."

In the frequency region around resonance the resistance change is fairly small and by itself would not affect the s.w.r. enough to matter, practically. The principal cause of the change in s.w.r. is the change in the reactive component of the antenna impedance when the frequency is varied. If the reactance changes rapidly with frequency the s.w.r. will rise rapidly off resonance, but if the rate of reactance change is small the shift in s.w.r. likewise will be small. Hence an antenna that has a relatively slow rate of reactance change will cover a wider frequency band, for a given value of s.w.r. at the band limits (such as 2 to 1 or 3 to 1), than one having a relatively rapid rate of reactance change.

Antenna Q

In the region around the resonant frequency of the antenna the impedance as measured at a current loop varies with frequency in essentially the same way as the impedance of a series-resonant circuit using lumped constants. It is therefore possible to define a quantity Q for the antenna in the same way as Q is defined in a series-resonant circuit. The Q of the antenna is a measure of the antenna's selectivity, just as the Q of an ordinary circuit is a measure of its selectivity.

The Q of the antenna can be found by measuring its input resistance and reactance at some frequency close to the resonant frequency (at exact resonance the antenna is purely resistive and there is no reactive component). Then, for frequency changes of less than 5 per cent from the exact resonant frequency, the antenna Q is given with sufficient accuracy by the following formula:

$$Q = \frac{X}{R} \cdot \frac{1}{2n}$$

where X and R are the measured reactance and resistance and n is the percentage difference, expressed as a decimal, between the antenna's resonant frequency and the frequency at which X and R were measured. For example, if the frequency used for the measurement differs from the resonant frequency by 2 per cent, $n =$ 0.02.

For an ordinary half-wave dipole over the range of length/diameter ratios used in Fig. 2-7, Chapter Two, the approximate Q values vary from about 14 for curve A to about 8 for curve C. In parasitic arrays with close spacing between elements the input Q may be well over 50, depending on the spacing and tuning (see Chapter Four).

S.W.R. vs. Q

If the Q of the antenna is known the variation in s.w.r. over the operating band can be determined from Fig. 3-82. It is assumed that the antenna is matched to the line at the center frequency of the band. Conversely, if a limit is set on the s.w.r., the width of the band that can be covered can be found from Fig. 3-82. As an example, suppose that a dipole having a Q of 15 (more or less typical of a wire antenna) is to be used over the 3.5-4 Mc. band and that it is

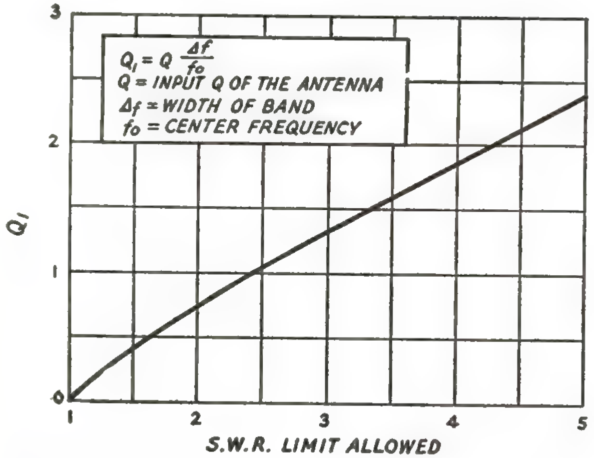


Fig. 3-82—Bandwidth in terms of s.w.r. limits, as a function of antenna Q. The inset formula gives an effective Q (Q_1 determined by the fractional band ($\Delta f/f_0$) and the actual antenna Q as defined in the text.

matched with a 1-to-1 s.w.r. at the band center. Then $\Delta f/f_0 = 0.5/3.75 = 0.133$ and $Q_1 = 15 \times 0.133 = 2$. The s.w.r. that can be expected at the band edges, 3.5 and 4 Mc., is shown by the chart to be a bit over 4 to 1. If it should be decided arbitrarily that no more than a 2-to-1 standing-wave ratio is allowable, Q_1 is 0.75 and from the formula in Fig. 3-82 the total bandwidth is found to be a little less than 200 kc.

Effect of Matching Network

The measurement of resistance and reactance to determine Q should be made at the input terminals of the matching network, if one is required. The selectivity of the matching network has just as much effect on the bandwidth, in terms of s.w.r. on the line, as the selectivity of the antenna itself. Where the greatest possible bandwidth is wanted a low-Q matching network must be used. This is not always controllable, particularly when the antenna resistance differs considerably from the Z_0 of the line to which it is to be matched. A large impedance ratio usually means that large values of reactance must be used in the matching section; in other words, the Q of the matching section in such cases tends to be higher than desirable. Simple systems having direct matching, such as a dipole fed with 75-ohm line or a folded dipole matched to the line, will have the greatest bandwidth, other things being equal, because no matching network is required.

Transmission-Line Measurements

The principal quantities to be measured on transmission lines are line current or voltage and the standing-wave ratio. Measurements of current or voltage are made for the purpose of determining the power input to the line. S.w.r. measurements are useful in connection with the design of coupling circuits and in the adjustment of the match between the antenna and transmission line, as well as in the adjustment of matching circuits as described earlier.

For most practical purposes a *relative* measurement is quite sufficient. An uncalibrated indicator that shows when the largest possible amount of power is being put into the line is just as useful, in most cases, as an instrument that measures the power accurately. It is seldom necessary to know the actual number of watts going into the line unless the overall efficiency of the system is being investigated. Again, an instrument that shows when the s.w.r. is close to 1 to 1 is all that is needed for impedance-matching adjustments. Accurate measurement of s.w.r. is necessary only in studies of antenna characteristics such as band width.

Quantitative measurements of reasonable accuracy demand good design and careful construction in the measuring instruments. They also require intelligent use of the equipment, including a knowledge not only of its limitations but also of stray effects that often lead to false results. A certain amount of skepticism regarding numerical data resulting from amateur measurements with simple equipment is justified until the complete conditions of the measurements are known. On the other hand, purely qualitative or relative measurements are easy to make and are quite reliable for the purposes mentioned above.

LINE CURRENT AND VOLTAGE

A current or voltage indicator that can be used with coaxial line is a useful piece of equipment. It need not be elaborate or expensive. Its principal function is to show when the maximum power is being taken from the transmitter; for any given set of line conditions (length, s.w.r., etc.) this will occur when the transmitter coupling is adjusted for maximum current or voltage at the input end of the line. Although the final-amplifier plate-current meter is frequently used for this purpose, it is not always a reliable indicator. In many cases, particularly with a screen-grid tube in the final stage, minimum loaded plate current does not occur simultaneously with maximum power output.

R.F. Voltmeter

A germanium diode in conjunction with a low-range milliammeter and a few resistors can be assembled to form an r.f. voltmeter suitable for connecting across the two conductors of a coaxial line, as shown in Fig. 3-83. It consists of a voltage divider R_1R_2 having a total resistance about 100 times the Z_0 of the line (so the power consumed will be negligible) with a

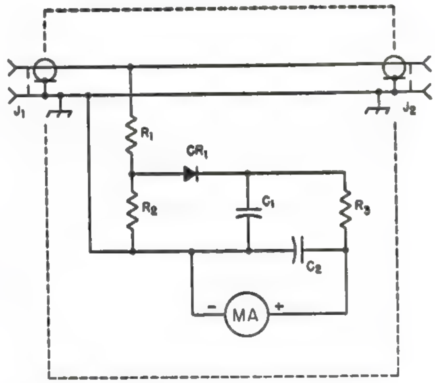


Fig. 3-83—R.f. voltmeter for coaxial line.

C_1, C_2—0.005 or 0.01 ceramic.

CR_1—Germanium crystal diode, any type.

J_1, J_2—Coaxial fittings, chassis-mounting type.

MA—0–1 milliammeter (more sensitive meter may be used if desired; see text).

R_1—6800 ohms, composition, 1 watt for each 100 watts of r.f. power.

R_2—680 ohms, ½ or 1 watt composition.

R_3—10,000 ohms ½ watt (see text).

diode rectifier and milliammeter connected across part of the divider to read relative r.f. voltage. The purpose of R_3 is to make the meter readings directly proportional to the applied voltage, as nearly as possible, by "swamping" the resistance of CR_1 since the latter resistance will vary with the amplitude of the current through the diode.

The voltmeter may be constructed in a small metal box, indicated by the dashed line in the drawing, fitted with coax receptacles. R_1 and R_2 should be composition resistors. The power rating for R_1 should be 1 watt for each 100 watts of carrier power in the matched line; separate 1- or 2-watt resistors should be used to make up the total power rating required, to a total resistance as given. Any type of resistor can be used for R_3; the total resistance should be such that about 10 volts d.c. will be developed across it at full scale. For example, a 0-1 milliammeter would require 10,000 ohms, a 0-500 microammeter would take 20,000 ohms, and so on. For a comparative measurements only, R_3 may be a variable resistor so the sensitivity can be adjusted for various power levels.

In constructing such a voltmeter care should be used to prevent inductive coupling between R_1 and the loop formed by R_2, CR_1 and C_1, and between the same loop and the line conductors in the assembly. With the lower end of R_1 dis-

connected from R_2 and grounded to the enclosure, but without changing its position with respect to the loop, there should be no meter indication when full power is going through the line.

If more than one resistor is used for R_1 the units should be arranged end to end with very short leads. R_1 and R_2 should be kept a half inch or more from metal surfaces parallel to the body of the resistor. If these precautions are observed the voltmeter will give consistent readings at frequencies up to 30 Mc. Stray capacitances and couplings limit the accuracy at higher frequencies but do not affect the utility of the instrument for comparative measurements.

Calibration

The meter may be calibrated in r.f. voltage by comparison with a standard such as an r.f. ammeter. This requires that the line be well matched so that the impedance at the point of measurement is equal to the actual Z_0 of the line. Since in that case $P = I^2Z_0$, the power can be calculated from the current. Then $E = \sqrt{PZ_0}$. By making current and voltage measurements at a number of different power levels, enough points may be secured to permit drawing a calibration curve for the voltmeter.

R.F. Ammeters

An r.f. ammeter can be mounted in any convenient location at the input end of the transmission line, the principal precaution in its mounting being that the capacitance to ground, chassis, and nearby conductors should be low. A bakelite-case instrument can be mounted on a metal panel without introducing enough shunt capacitance to ground to cause serious error up to 30 Mc. When installing a metal-case instrument on a metal panel it should be mounted on a separate sheet of insulating material in such a way that there is an eighth of an inch or more separation between the edge of the case and the metal.

Fig. 3-84—A convenient method of mounting an r.f. ammeter for use in coaxial line. This is a metal-case instrument mounted on a thin bakelite panel, the diameter of the cut-out in the metal being such as to clear the edge of the meter by about an eighth inch.

A two-inch instrument can be mounted in a $2 \times 4 \times 4$ inch metal box as shown in Fig. 3-84. This is a convenient arrangement for use with coaxial line.

Installed as just described, a good-quality r.f. ammeter will measure current with an accuracy that is entirely adequate for calculating power in the line. As discussed above in connection with calibrating r.f. voltmeters, the line must be closely matched by its load so the actual impedance will be resistive and equal to Z_0. The scales of such instruments are cramped at the low end, however, which limits the range of power that can be measured by a single meter. The useful current range is about 3 to 1, corresponding to a power range of about 9 to 1.

For additional details on the use of r.f. current indicators, see discussion in the earlier section on impedance-matching or coupling circuits.

S.W.R. MEASUREMENTS

On parallel-conductor lines it is possible to measure the standing-wave ratio by moving a current (or voltage) indicator along the line, noting the maximum and minimum values of current (or voltage) and then computing the s.w.r. from these measured values. This cannot be done with coaxial line since it is not possible to make measurements of this type inside the cable. The technique is in fact seldom used with open lines, because it is not only inconvenient but sometimes impossible to reach all parts of the line conductors. Also, the method is subject to considerable error from antenna currents flowing on the line.

Present-day s.w.r. measurements are practically always made with some form of "directional coupler" or r.f. bridge circuit. The bridges themselves are fundamentally simple, but considerable care is required in their construction if the measurements are to be accurate. The requirements for bridges used only for the adjustment of impedance-matching circuits, rather than actual s.w.r. measurement, are not so stringent and a bridge for this purpose can be made quite easily.

Bridge Circuits

Three commonly-used bridge circuits are shown in Fig. 3-85. The bridges consist essentially of two voltage dividers in parallel, with a voltmeter connected between the junctions of each pair of "arms", as the individual elements are called. When the equations shown to the right of each circuit are satisfied there is no potential difference between the two junctions

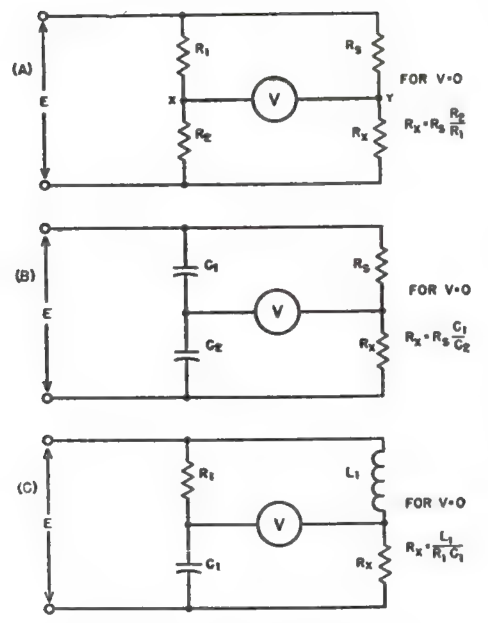


Fig. 3-85—Bridge circuits suitable for s.w.r. measurement. A—Wheatstone type using resistance arms. B—Capacitance-resistance bridge ("Micromatch"). C—Maxwell bridge. Conditions for balance are independent of frequency in all three types.

and the voltmeter indicates zero voltage. The bridge is then said to be in "balance".

Taking Fig. 3-85A as an illustration, if $R_1 = R_2$ half the applied voltage, E, will appear across each resistor. Then if $R_s = R_x$, ½ E will appear across each of these resistors and the voltmeter reading will be zero. Remembering that a matched transmission line has a purely-resistive input impedance, suppose that the input terminals of such a line are substituted for R_x. Then if R_s is a resistor equal to the Z_0 of the line the bridge will be balanced. If the line is not perfectly matched, its input impedance will not equal Z_0 and hence will not equal R_s since the latter is chosen equal to Z_0. There will then be a difference if potential between points X and Y and the voltmeter will show a reading. Such a bridge therefore can be used to show the presence of standing waves on the line, because the line input impedance will be equal to Z_0 only when there are no standing waves.

Considering the nature of the incident and reflected components of voltage that make up the actual voltage at the input terminals of the line as discussed earlier, it should be clear that when $R_s = Z_0$, the bridge is always in balance for the incident component. Thus the voltmeter does not respond to the incident component at any time but reads only the reflected component (assuming that R_2 is very small compared with the voltmeter impedance). The incident component can be measured across either R_1 or R_2, if they are equal resistances. The standing wave ratio is then where E_1 is the incident voltage and E_2 is the reflected voltage. It is

$$S.W.R. = \frac{E_1 + E_2}{E_1 - E_2}$$

often simpler to normalize the voltages by expressing E_2 as a fraction of E_1, in which case the formula becomes

$$S.W.R. = \frac{1 + k}{1 - k}$$

where $k = E_2/E_1$.

The operation of the other two circuits in Fig. 3-85 is essentially the same, although these two circuits have arms containing reactance as well as resistance.

It is not necessary that $R_1 = R_2$ in Fig. 3-85A; the bridge can be balanced, in theory, with any ratio of these two resistances provided R_s is changed accordingly. However, the accuracy is highest, in practice, when the two are equal, and this circuit is generally so used. In the circuit at B the bridge "ratio" is usually quite high—that is, R_s is of the order of 1/50 the resistance of R_x, the Z_0 of the line with which the bridge is to be used. The reason for this is that if R_s is of the order of one ohm the bridge can be used at quite high power levels since the power lost in R_s is small and no power is consumed in C_1 or C_2. However, a high bridge ratio makes construction for accurate measurement much more difficult.

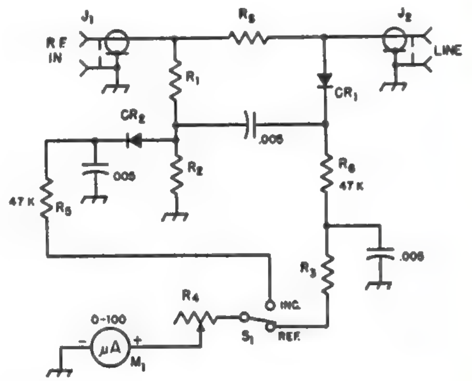


Fig. 3-86—Resistance bridge for s.w.r. measurement. Capacitors are disk ceramic. Resistors are ½ watt composition except as noted below.

CR_1, CR_2—Germanium diode, high back-resistance type (1N54A, etc.).

J_1, J_2—Coaxial connectors, chassis-mounting type.

M_1—0–100 d.c. microammeter.

R_1, R_2—47 ohms, ½ watt composition (see text).

R_3—See text.

R_4—50,000-ohm volume control.

R_s—Resistance equal to line Z_0 (½ or 1 watt composition).

S_1—S.p.d.t. toggle.

Resistance Bridge

The basic bridge type shown in Fig. 3-85A is recommended for home construction if the bridge is to be used for actual s.w.r. measurement. A practical circuit for such a bridge is given in Fig. 3-86, and a representative layout is shown in Fig. 3-87. Properly built, a bridge of this design can be used for measurement of standing-wave ratios up to about 15 to 1 with good accuracy.

Important constructional points to be observed are:

1) Keep leads in the r.f. circuit short, to reduce stray inductance.

2) Mount resistors two or three times their body diameter away from metal parts, to reduce stray capacitance.

3) Place the r.f. components so there is as little inductive and capacitive coupling as possible between the bridge arms.

In the layout shown in Fig. 3-87 the input and line connectors, J_1 and J_2, are mounted fairly close together so the standard resistor, R_s, can be supported with short leads directly between the center terminals of the connectors. R_2 is mounted at right angles to R_s, and a shield partition is used between these two components and the others.

The two 47,000-ohm resistors, R_5 and R_6, in Fig. 3-86 are voltmeter multipliers for the 0-100 microammeter used as an indicator. This is sufficient resistance to make the voltmeter linear (that is, the meter reading is directly proportional to the r.f. voltage) and no voltage calibration curve is needed. CR_1 is the rectifier for the reflected voltage and CR_2 is for the incident voltage. Because of resistor tolerances and small differences in diodes, the readings may differ slightly with two multipliers of the same nominal resistance, so a correction resistor R_3 is included in the circuit. Its value should be selected so that the meter reading is the same with S_1 in either position, when r.f. is applied to the bridge with the line connection open. In the instrument shown, a value of 1000 ohms was required in series with the multiplier for reflected voltage; in other cases different values probably would be needed and R_3 might have to be put in series with the multiplier for the incident voltage. This can be determined by experiment.

The value used for R_1 and R_2 is not critical, but the two resistors should be matched to within 1 or 2 per cent if possible. The resistance of R_s should be as close as possible to the actual Z_0 of the line to be used (generally 52 or 75 ohms). The resistor should be selected by actual measurement with an accurate resistance bridge if one is available.

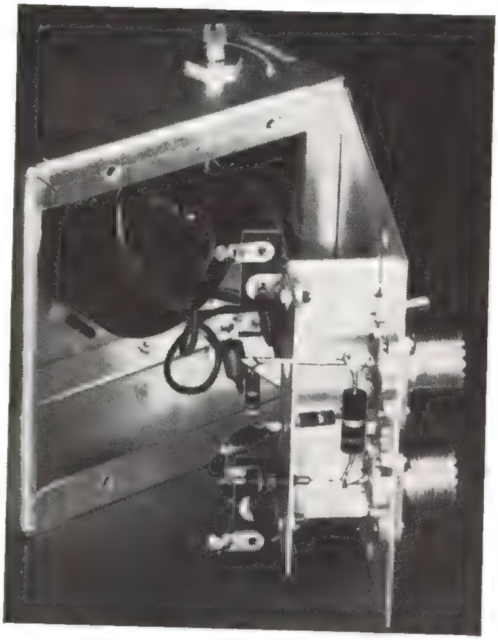


Fig. 3-87—A 2 × 4 × 4 inch aluminum box is used to house this s.w.r. bridge, which uses the circuit of Fig. 3-86. The variable resistor R_4 is mounted on the side.

The bridge components are mounted on one side plate of the box and a miniature chassis formed from a piece of aluminum. The input connector is at the top in this view. R_s is connected directly between the two center posts of the connectors. R_2 is visible behind it and perpendicular to it. One terminal of CR_1 projects through a hole in the chassis so the lead can be connected to J_2. R_1 is mounted vertically to the left of the chassis in this view, with CR_2 connected between the junction of R_1R_2 and a tie point.

R_4 is for adjusting the incident-voltage reading to full scale in the measurement procedure described below. Its use is not essential, but it offers a convenient alternative to exact adjustment of the r.f. input voltage.

Testing

R_1, R_2, and R_s should be measured with a good ohmmeter or resistance bridge after wiring is completed, in order to make sure their values have not changed from the heat of soldering. Disconnect one side of the microammeter and leave the input and output terminals of the unit open during such measurements, in order to avoid stray shunt paths through the rectifiers.

Check the two voltmeter circuits as described above, applying enough r.f. (about 10 volts) to

the input terminals to give a full-scale reading with the line terminals open. If necessary, try different values for R_3 until the reading is the same with S_1 in either position.

With J_2 open, adjust the r.f. input voltage and/or R_4 for full-scale reading with S_1 in the reflected-voltage position. Then short-circuit J_2 by touching a screwdriver between the center terminal and the frame of the connector to make a low-inductance short. Throw S_1 to the incident-voltage position and readjust R_4 for full scale, if necessary. Then throw S_1 to the reflected-voltage position, keeping J_2 shorted, and the reading should be full scale as before. If it is not, R_1 and R_2 are not the same value, or there is stray coupling between the arms of the bridge. It is necessary that the reflected voltage read full scale with J_2 either open or shorted, when the incident voltage is set to full scale in each case, in order to make accurate s.w.r. measurements.

The circuit should pass these tests at all frequencies at which it is to be used. It is sufficient to test at the lowest and highest frequencies, usually 3.5 and 28 or 50 Mc. If R_1 and R_2 are poorly matched but the bridge construction is otherwise good, discrepancies in the readings will be substantially the same at all frequencies. A difference in behavior at the low and high ends of the frequency range can be attributed to stray couplings between bridge arms, or stray inductance or capacitance in the arms.

To check the bridge for balance, apply r.f. and adjust R_4 for full scale with J_2 open. Then connect a resistor identical with R_s (the resistance should match within 1 or 2 per cent) to the line terminals, using the shortest possible leads. It is convenient to mount the test resistor inside a cable connector (PL-259), a method of mounting that also minimizes lead inductance. When the test resistor is connected, the reflected-voltage reading should drop to zero. The incident voltage should be reset to full scale by means of R_4, if necessary. The reading should be zero at any frequency in the range to be used. If a good null is obtained at low frequencies but some residual current shows at the high end, the trouble may be the inductance of the test resistor leads, although it may also be caused by stray couplings between the arms of the bridge itself. If there is a constant low (but not zero) reading at all frequencies the cause is poor matching of the resistance values. Both effects can be present simultaneously. A good null must be obtained at all frequencies before the bridge is ready for use.

Reducing Power for Bridge Operation

The r.f. power input to a bridge of this type must be limited to a few watts at most, because of the power-dissipation ratings of the resistors. If the transmitter has no provision for reducing power output to a very low value—less than 5 watts—a simple "power absorber" circuit can be made up as shown in Fig. 3-88. The lamp I_1 tends to maintain constant current through the resistor over a fairly wide power range, so the voltage drop across the resistor also tends to be constant. This voltage is applied to the bridge, and with the constants given is in the right range for resistance-type bridges.

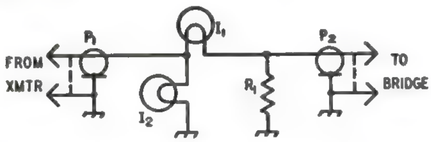


Fig. 3-88—"Power absorber" circuit for use with resistance-type s.w.r. bridges when the transmitter has no special provisions for power reduction. For r.f. powers up to 50 watts, I_1 is a 115-volt 40 watt incandescent lamp and I_2 is not used. For higher powers use sufficient additional lamp capacity at I_2 to load the transmitter about to normal output; for example, for 250 watts output I_2 may consist of two 100-watt lamps in parallel. R_1 is made from three 1-watt 68-ohm resistors connected in parallel. P_1 and P_2 are cable-mounting coaxial connectors.

Leads in the circuit formed by the lamps and R_1 should be kept short, but convenient lengths of cable may be used between this assembly and the connectors.

Measuring S.W.R.

To make a measurement, connect the line to J_2 and apply sufficient r.f. voltage to J_1 to give a full-scale incident-voltage reading. R_4 may be used to set to exactly full scale. Then throw S_1 to the reflected-voltage position and note the meter reading. The s.w.r. is then found by substituting the readings in the formula previously given.

For example, if the full-scale calibration of the d.c. instrument is 100 microamperes and the reading with S_1 in the reflected-voltage position is 40 microamperes, the s.w.r. is

$$S.W.R. = \frac{100+40}{100-40} = \frac{140}{60} = 2.34 \text{ to } 1.$$

The meter scale may be calibrated in any arbitrary units so long as the scale has equal divisions, since it is the ratios of the voltages and not the actual values that determine the s.w.r.

AVOIDING ERRORS IN S.W.R. MEASUREMENTS

The principal causes of inaccuracies within the bridge are differences in the resistances of R_1 and R_2, stray inductance and capacitance in the bridge arms, and stray coupling between arms. If the checking procedure described above is followed through carefully, the bridge of Fig. 3-86 should be amply accurate for practical use. The accuracy is highest for low standing-wave ratios because of the nature of the s.w.r. calculation; at high ratios the divisor in the equation above represents the difference between two nearly equal quantities, so a small

error in voltage measurement may mean a considerable difference in the calculated s.w.r.

Voltmeter Impedance and Linearity

It is important that the voltmeter have high impedance. With the series diode arrangement used in Fig. 3-86 the r.f. impedance is approximately equal to one-half the d.c. resistance in the meter circuit, and so the impedance is 20,000 ohms or more in this case. This is high enough to have negligible effect on the impedances in the bridge arms. It is also desirable that the voltmeter be linear since this avoids the necessity for a special calibration. Adequate linearity requires that the d.c. resistance in the meter circuit be in the thousands of ohms, which in turn requires the use of a relatively sensitive d.c. instrument.

Incident Voltage

It is essential that the incident voltage be measured, since it will usually change when a load is connected to the bridge because of the change in the load on the source of r.f. Measurements made with simple bridges in which no provision is made for checking the incident voltage under load are of little value except in the special case of a 1 to 1 s.w.r., where the reflected voltage is zero for any value of incident voltage.

Standard Resistor

The standard resistor R_s must equal the actual Z_0 of the line. The actual Z_0 of a sample of line may differ by a few percent from the nominal figure because of manufacturing variations, but this has to be tolerated. In the 50- to 75-ohm range, the r.f. resistance of a composition resistor of ½ or 1 watt rating is essentially identical with its d.c. resistance.

"Antenna" Currents

As explained earlier, there are two ways in which "parallel" or "antenna" currents can be caused to flow on the outside of a coaxial line —currents induced on the line because of its spatial relationship to the antenna (Fig. 3-48) and currents that result from the direct connection between the coax outer conductor and (usually) one side of the antenna. The induced current usually will not be troublesome if the bridge and the transmitter (or other source of r.f. power for operating the bridge) are shielded so that any r.f. currents flowing on the *outside* of the line cannot find their way into the bridge. This point can be checked by cutting in an additional section of line (⅛ to ¼ electrical wavelength, preferably) of the same Z_0. The s.w.r. indicated by the bridge should not change except possibly for a slight decrease because of the additional line loss, as discussed earlier in this chapter. If there is a marked change better shielding may be required.

Parallel-type currents caused by the connection to the antenna will cause a change in s.w.r. with line length even though the bridge and transmitter are well shielded and the shielding is maintained throughout the system by the use of coaxial fittings. This is because the outside of the coax tends to become part of the antenna system, being connected to the antenna at the feed point, and so constitutes a load on the line along with the desired load represented by the antenna itself. The s.w.r. on the line then is determined by the composite load of the antenna and the outside of the coax, and since changing the line length changes one component of this composite load, the s.w.r. changes too.

The remedy for such a situation is to use a good balun or to detune the outside of the line by proper choice of length. It is well to note that this is not a measurement error, since what the instrument reads is the actual s.w.r. on the line. However, it is an undesirable condition since the line is operating at a higher s.w.r. than it should—and would—if the parallel-type current on the outside of the coax were eliminated.

Spurious Frequencies

Off-frequency components in the r.f. voltage applied to the bridge may cause considerable error. The principal components of this type are harmonics and low-frequency "subharmonics" that may be fed through the final stage of the transmitter driving the bridge. The antenna is almost always a fairly selective circuit, and even though the system may be operating with a very low s.w.r. at the desired frequency it is practically always mismatched at harmonic and subharmonic frequencies. If such spurious frequencies are applied to the bridge in appreciable amplitude the s.w.r. indication will be very much in error. In particular, it may not be possible to secure a null on the bridge voltmeter with any set of adjustments of the matching circuit. The only remedy is to filter out the unwanted components by increasing the selectivity of the circuits between the transmitter's final amplifier and the bridge.

Parallel-Conductor Lines

Measurements made directly on parallel-con-

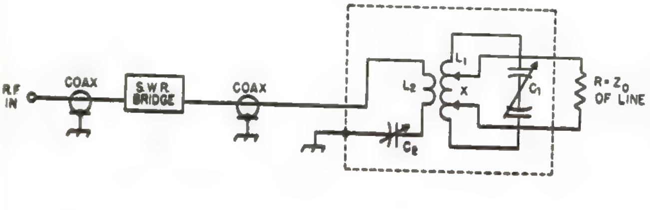


Fig. 3-89—Initial setup for s.w.r. measurement in parallel-conductor line. Actual measurement is made with a coax-type bridge in coaxial line, through a matching circuit of the type used for coupling between coax and parallel-conductor line.

ductor lines are not generally reliable because there are nearly always "antenna currents" present to a greater or lesser degree on a transmission line. With ordinary parallel-conductor lines it is not possible to separate the "transmission-line" and "antenna" currents. The latter therefore can actuate the bridge voltmeter and may cause considerable error.

The effect of such parallel currents can be reduced and usually made negligible by the use of a suitable coupling circuit or balun between the line and a coaxial-type s.w.r. bridge. The inductively-coupled matching circuit described in detail earlier is quite suitable and can be used between coaxial line and a parallel-conductor line of any Z_0.

If the coupling coil L_2 in Fig. 3-89 is coupled to the center of L_1 so that there is equal coupling to both halves of L_1, and if the capacitive coupling between the two coils is small, then parallel or antenna currents flowing in L_1 induce equal and opposite voltages in L_2. The resultant voltage caused by such currents is zero, consequently. On the other hand, the reflected component of current caused by a mismatch between the antenna and line is a transmission-line current and will induce a voltage in L_2 in the normal way. If the circuit is set up to transform the Z_0 of the parallel-conductor line to that of the coaxial line, and if the losses in the matching circuit are small (which is usually the case) the voltage induced in L_2C_2 will have the same relationship to the incident voltage in that circuit as exists between reflected and incident voltages on the parallel-conductor line. Hence the s.w.r. can be measured in the coaxial line just as readily as in the parallel-conductor line itself, and without the presence of antenna currents to cause error.

The test setup is shown in Fig. 3-89. The constants of the matching circuit are the same as discussed earlier in this chapter. The line should be disconnected from the circuit and a non-inductive resistor (composition type, 1-watt rating satisfactory) substituted. Its resistance should be equal to the Z_0 of the type of line used—300 ohms for Twin-Lead, and so on. The circuit should be adjusted to give a 1-to-1 s.w.r. in the coax line, after which the resistor should be removed and the line reconnected to the same taps on L_1 as the test resistor, *without touching any of the matching-circuit adjustments.* Then the measured s.w.r. in the coax is also the s.w.r. in the parallel-conductor line.

Grounding point X, the center of L_1, will reduce capacitive coupling between L_1 and L_2, and is desirable in those cases where the length of the parallel-conductor line is such as to bring a voltage loop at or near its input end. If the line length is such as to bring a current loop near the input end it is better to leave the center of L_1 ungrounded, since antenna currents will not flow through the coil with the center tap open except as a result of unbalanced capacitances to ground. The effects of such unbalances should be small at a current loop.

Bridges for Matching

Considerable attention was devoted to the resistance-type s.w.r. bridge in the preceding section because it is the simplest type that is capable of adequate accuracy in measuring voltage standing-wave ratio. Its disadvantage is that it must be operated at a very low power level, and thus is not suitable for continuous monitoring of the s.w.r. in actual transmission. To do this the bridge or reflectometer must be capable of carrying the entire power output of the transmitter, and should do it with negligible loss.

A number of reflectometers are available commercially that can be left permanently in the transmission line. Some of these are calibrated in forward and reflected watts (usually the calibration is for 50-ohm line, and does not hold for other values of line Z_0). Types with reasonably-accurate power calibrations are relatively expensive, but if one of these is used the reflected/forward power ratio can easily be converted into the corresponding voltage ratio for determining v.s.w.r. by the formula given earlier. Since power is proportional to voltage squared, the normalized formula becomes

$$V.S.W.R. = \frac{1 + \sqrt{k}}{1 - \sqrt{k}}$$

where k is the ratio of reflected power to forward power. A chart for determining v.s.w.r. from either voltage or power measurements is given in Fig. 3-90.

Low-cost reflectometers which do not have a guaranteed wattmeter calibration are not ordinarily reliable for numerical measurement of

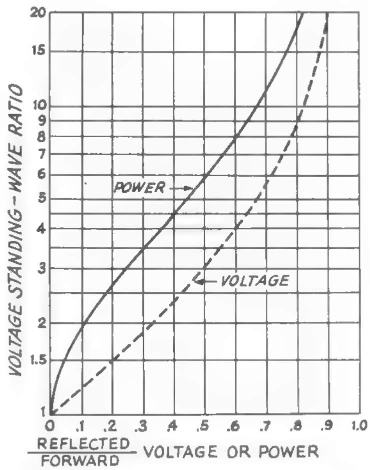


Fig. 3-90—Chart for finding voltage standing-wave ratio when the ratio of reflected-to-forward voltage or reflected-to-forward power is known.

standing-wave ratio. They are, however, very useful as aids in the adjustment of matching networks, since the object in such adjustment is to reduce the reflected voltage or power to zero. Relatively-inexpensive devices can be used for this, since only good bridge balance is required, not actual calibration. Bridges of this type are usually "frequency-sensitive"—that is, the meter response becomes greater with increasing frequency, for the same applied voltage. When matching and line monitoring, rather than s.w.r. measurement, is the principal use of the bridge this is not a serious handicap.

Various simple reflectometers useful for matching and monitoring have been described from time to time in *QST* and in *The Radio Amateur's Handbook.* Because most of these are frequency-sensitive it is difficult to calibrate them accurately for s.w.r. measurement, but their low cost and suitability for use at any power level, combined with the ability to show accurately when a matching circuit has been properly adjusted, make them a worth-while addition to the amateur station. The latest edition of the *Handbook* should be consulted for the most modern circuits.

Multielement Directive Arrays

The gain and directivity that can be secured by intentionally combining antenna elements into an array represent a worth-while improvement both in transmitting and receiving. Power gain in an antenna is the same as an equivalent increase in the transmitter power. But, unlike increasing the power of one's own transmitter, it works equally well on signals received from the favored direction. In addition, the directivity reduces the strength of signals coming from the directions not favored, and so helps discriminate against a great deal of interference.

One common method of securing gain and directivity is to combine the radiation from a group of half-wave dipoles in such a way as to concentrate it in a desired direction. The way in which such combinations affect the directivity has been explained in Chapter Two. A few words of additional explanation may help make it clear how power gain is achieved.

In Fig. 4-1, imagine that the four circles, *A, B, C, D,* represent four dipoles so far separated from each other that the coupling between them is negligible. The point *P* is supposed to be so far away from the dipoles that the distance from *P* to each one is exactly the same (obviously *P* would have to be much farther away than it is in this drawing). Under these conditions the fields from all the dipoles will add up at *P*, if all four are fed r.f. currents in the same phase.

Let us say that a certain current, *I*, in dipole *A* will produce a certain value of field strength, *E*, at the distant point *P*. The same current in any of the other dipoles will produce the same field at *P*. Thus if only dipoles *A* and *B* are operating, each with a current *I*, the field at *P* will be 2*E*.

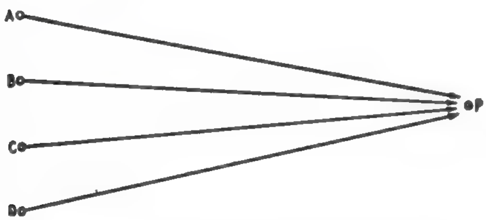


Fig. 4-1—Fields frpm separate antennas combine at a distant point, P, to produce a field strength that exceeds the field produced by the same power in a single antenna.

With *A, B,* and *C* operating the field will be 3*E*, and with all four operating with the same *I*, the field will be 4*E*. Since the power received at *P* is proportional to the square of the field strength, the relative power received at *P* is 1, 4, 9 and 16, depending on whether one, two, three or four dipoles are operating.

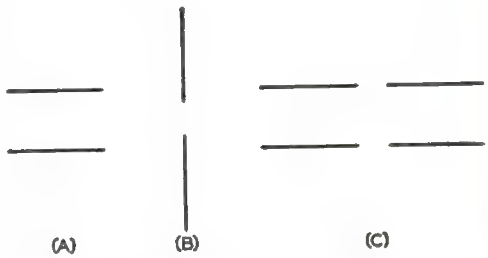


Fig. 4-2—Parallel (A) and collinear (B) antenna elements. The array shown at C combines both parallel and collinear elements.

Now since all four dipoles are alike and there is no coupling between them, the same power must be put into each in order to cause the current *I* to flow. For two dipoles the relative power input is 2, for three dipoles it is 3, for four dipoles 4, and so on. The gain in each case is the relative received (or output) power divided by the relative input power. Thus we have:

Dipoles	*Relative Output Power*	*Relative Input Power*	*Power Gain*	*Gain in Db.*
A only	1	1	1	0
A and *B*	4	2	2	3
A, B and *C*	9	3	3	4.8
A, B, C and *D*	16	4	4	6

The power gain is directly proportional to the number of elements used.

It is well to have clearly in mind the conditions under which this relationship is true:

1) The fields from the separate antenna elements must be in phase at the receiving point.

2) The currents in all elements must be identical.

3) The elements must be separated in such a way that the current induced in one by another is negligible; i.e., the radiation resistance of each element must be the same as it would have been had the other elements not been there.

Very few antenna arrays meet all these conditions exactly. However, as a rough approximation it may be said that the power gain of a directive array consisting of dipole elements in which optimum values of element spacing are used is proportional to the number of elements. It is not impossible, though, for an estimate based on this rule to be in error by a factor of 2 or more.

Definitions

The "element" in a multielement directive array is usually a half-wave dipole. The length is not always an exact electrical half wavelength, because in some types of arrays it is desirable that the element show either inductive or capacitive reactance. However, the departure in length from a true half wave is ordinarily small (not more than 5%, in the usual case) and so has no appreciable effect on the radiating properties of the element.

Antenna elements in multielement arrays of the type considered in this chapter are always

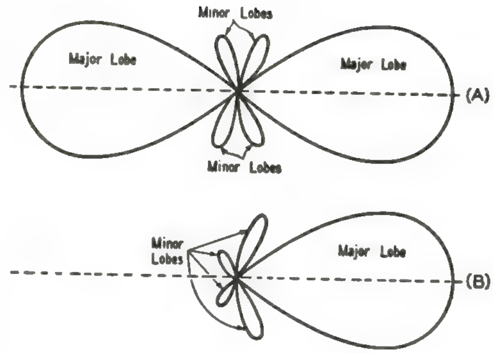


Fig. 4-4—Typical bidirectional (A) and unidirectional (B) directive patterns. These drawings also illustrate the application of the terms "major" and "minor" to the pattern lobes.

either parallel, as at A in Fig. 4-2, or collinear (end-to-end), Fig. 4-2B. Fig. 4-2C shows an array combining both parallel and collinear elements. The elements can be either horizontal or vertical, depending on whether horizontal or vertical polarization is desired. There is seldom any reason for mixing polarization, so arrays are customarily constructed with all elements similarly polarized.

A **driven element** is one supplied power from the transmitter, usually through a transmission line. A **parasitic element** is one that obtains power solely through coupling to another element in the array because of its proximity to such an element.

A **driven array** is one in which all the elements are driven elements. A **parasitic array** is one in which one or more of the elements are parasitic elements. At least one element in a parasitic array has to be a driven element, since it is necessary to introduce power into the array.

A **broadside array** is one in which the principal direction of radiation is perpendicular to the axis of the array and to the plane containing the elements. An **end-fire array** is one in which the principal direction of radiation coincides with the direction of the array axis. These definitions are illustrated by Fig. 4-3.

A **bidirectional array** is one that radiates equally well in either direction along the line of maximum radiation. A bidirectional pattern is shown in Fig. 4-4 at A. A **unidirectional array** is one that has only one principal direction of radiation, as illustrated by the pattern at B in Fig. 4-4.

The **major lobes** of the directive pattern are those in which the radiation is maximum. Lobes of lesser radiation intensity are called **minor lobes.** The **beamwidth** of a directive antenna is the width, in degrees, of the major lobe between the two directions at which the relative radiated power is equal to one-half its value at the peak of the lobe. At these "half-power points" the field intensity is equal to 0.707 times its maxi-

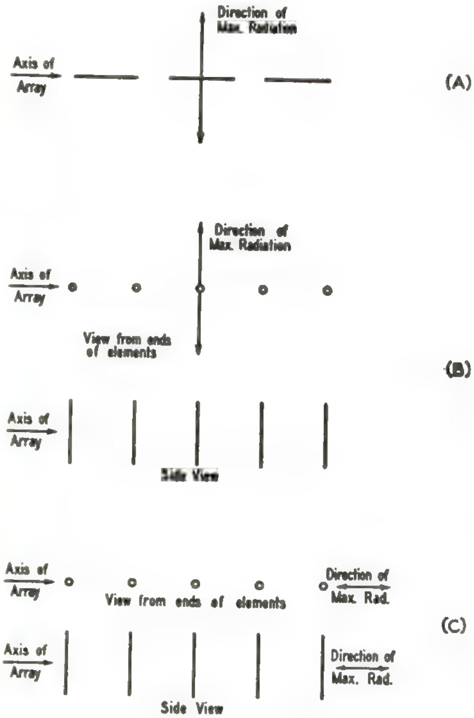


Fig. 4-3—Representative broadside arrays are shown at A and B, the first with collinear elements, the second with parallel elements. An end-fire array is shown at C. Practical arrays may combine both broadside and end-fire directivitly, including both parallel and collinear elements.

mum value, or down 3 db. from maximum. Fig. 4-5 is an example of a lobe having a beam width of 30 degrees.

Unless specified otherwise, the term "gain" as used in this chapter is the power gain over a half-wave dipole of the same orientation and height as the array under discussion, and having the same power input. Gain may either be measured experimentally or determined by calculation. Experimental measurement is difficult and often subject to considerable error, since in addition to the normal errors in measurement (the accuracy of simple r.f. measuring equipment is relatively poor, and even high-quality instruments suffer in accuracy compared with their low-frequency and d.c. counterparts) the accuracy depends considerably on the conditions—the antenna site, including height, terrain characteristics, and surroundings—under which the measurements are made. Calculations are frequently based on the measured or theoretical directive patterns (see Chapter Two) of the antenna. An approximate formula often used is

$$G = \frac{30{,}000}{\theta_H \theta_V}$$

where

G = gain over an *isotropic* antenna.

θ_H = horizontal half-power beam width in degrees.

θ_V = vertical half-power beam width in degrees.

This formula, strictly speaking, applies only to antennas having approximately equal and narrow *E*- and *H*-plane (see Chapter Two) beamwidths—up to about 20 degrees—and no large minor lobes. The error may be considerable when the formula is applied to simple directive antennas having relatively large beamwidths. The error is in the direction of making the calculated gain larger than the actual gain.

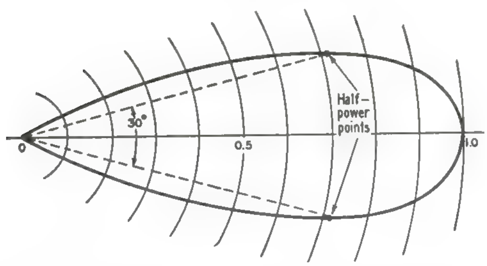


Fig. 4-5—The width of a beam is the angular distance between the directions at which the receiver or transmitted power is one-half the maximum power.

Front-to-back ratio means the ratio of the power radiated in the favored direction to the power radiated in the opposite direction.

Phase

The term "phase" has the same meaning when used in connection with the currents flowing in antenna elements as it does in ordinary circuit work. For example, two currents are in phase when they reach their maximum values, flowing in the same direction, at the same instant. The direction of current flow depends on the way in which power is applied to the element.

This is illustrated in Fig. 4-6. Assume that by some means an identical voltage is applied to each of the dipoles at the end marked A. Assume also that the instantaneous polarity of the voltage is such that the current is flowing away from the point at which the voltage is applied. The arrows show the assumed current directions. Then the currents in elements *1* and *2* are completely in phase, since they are flowing in the same direction in space and are caused by the same voltage. However, the current in element *3* is flowing in the *opposite* direction in space because the voltage is applied to the opposite end of the element. The current in element *3* is therefore 180 degrees out of phase with the currents in elements *1* and *2*.

The phasing of driven elements depends on the direction of the element, the phase of the applied voltage, and the point at which the voltage is applied. In the systems used by amateurs the voltages applied to the elements are practically always exactly in or exactly out of phase with each other. Also, the axes of the elements are always in the same direction, since parallel or collinear elements are invariably used. The currents in driven elements in such systems are therefore always either exactly in or out of phase with the currents in other elements.

It is possible to use phase differences of less than 180 degrees in driven arrays—one important case is where the voltage applied to one set of elements differs by 90 degrees from the voltage applied to another set—but such systems have not met with much application in amateur work. The reason probably is that making provision for proper phasing is considerably more of a problem than in the case of simple 0- or 180- degree phasing.

In parasitic arrays the phase of the currents in the parasitic elements depends on the spacing and tuning, as described later.

Ground Effects

The effect of the ground is the same with a directive antenna as it is with a simple dipole antenna. The reflection factors discussed in Chapter Two may therefore be applied to the vertical

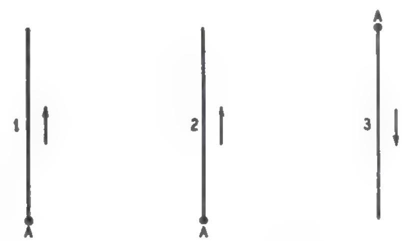


Fig. 4-6—Illustrating phasing of currents in antenna elements.

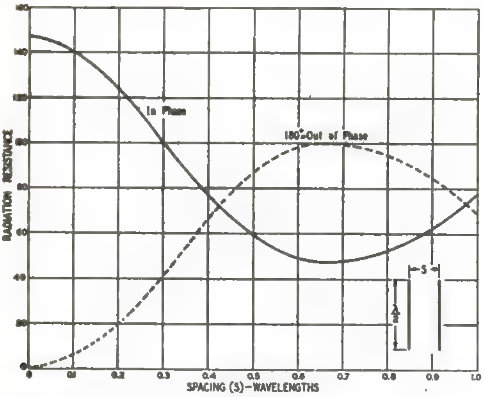


Fig. 4-7—Radiation resistance measured at the center of one element as a function of the spacing between two parallel half-wave self-resonant antenna elements.

pattern of an array, subject to the same modifications mentioned in that chapter. In cases where the array elements are not all at the same height, the reflection factor for the *mean* height of the array must be used. The mean height is the average of the heights measured from the ground to the centers of the lowest and highest elements.

MUTUAL IMPEDANCE

Consider two half-wave dipoles that are fairly close to each other. When power is applied to one and current flows, a voltage will be induced in the second by the electromagnetic field and current will flow in it as well. The current in antenna No. 2 will in turn induce a voltage in antenna No. 1, causing a current to flow in the latter. The total current in No. 1 is then the sum (taking phase into account) of the original current and the induced current.

If the voltage applied to antenna No. 1 has not changed, the fact that the amplitude of the current flowing is different, with antenna No. 2 present, than it would have been had No. 2 not been there indicates that the presence of the second antenna has changed the impedance of the first. This coupling effect is called **mutual impedance.** The actual impedance of an antenna element is the sum of its self-impedance (the impedance with no other antennas present) and its mutual impedance with all other antennas in the vicinity.

The magnitude and nature of the mutual impedance depends on the amplitude of the current induced in the first antenna by the second, and on the phase relationship between the original and induced currents. The amplitude and phase of the induced current depend on the spacing between the antennas and whether or not the second antenna is tuned to resonance.

Amplitude of Induced Current

The induced current will be largest when the two antennas are close together and are parallel. Under these conditions the voltage induced in the second antenna by the first, and in the first by the second, has its greatest value and causes the largest current flow. The coupling decreases as the parallel antennas are moved farther apart.

The coupling between collinear antennas is comparatively small, and so the mutual impedance between such antennas is likewise small. It is not negligible, however.

Phase Relationships

When the separation between the two antennas is an appreciable fraction of a wavelength a measurable period of time elapses before the field from antenna No. 1 reaches antenna No. 2 and there is a similar time lapse before the field set up by the current in No. 2 gets back to induce a current in No. 1. Hence the current induced in No. 1 by No. 2 will have a phase relationship with the original current in No. 1 that depends on the spacing between the two antennas.

The induced current can range all the way from being completely in phase with the original current to being completely out of phase with it. In the first case the total current is larger than the original current and the antenna impedance is reduced. In the second, the total current is smaller and the impedance is increased. At intermediate phase relationships the impedance will be lowered or raised depending on whether the induced current is mostly in or mostly out of phase with the original current.

Except in the special cases when the induced current is exactly in or out of phase with the original current, the induced current causes the phase of the total current to shift with respect to the applied voltage. The mutual impedance, in other words, has both resistive and reactive components. Consequently, the presence of a second antenna nearby may cause the impedance of an antenna to be reactive—that is, the antenna will be detuned from resonance—even though its self-impedance is entirely resistive. The amount of detuning depends on the magnitude and phase of the induced current.

Tuning Conditions

A third factor that affects the impedance of antenna No. 1 when No. 2 is present is the tuning of the latter. If No. 2 is not exactly resonant the current that flows in it as a result of the induced voltage will either lead or lag the phase it would have had if the antenna were resonant. This causes an additional phase advance or delay that affects the phase of the current induced back in No. 1. Such a phase lag has an effect similar to a change in the spacing between self-resonant antennas. However, a change in tuning is not exactly equivalent to a change in spacing because the two methods do not have the same effect on the amplitude of the induced current.

Mutual Impedance and Gain

The mutual impedance between antennas is important because it determines the amount of current that will flow for a given amount of power supplied. It must be remembered that it is the *current* that determines the field strength

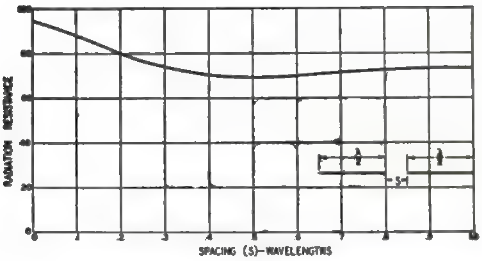


Fig. 4-8—Radiation resistance measured at the center of one element as a function of the spacing between the ends of two collinear self-resonant half-wave antenna elements operated in phase.

from the antenna. Other things being equal, if the mutual impedance between two antennas is such that the currents are greater, for the same total power, than would be the case if the two antennas were not coupled, the power gain will be greater than in the case discussed at the beginning of this chapter. On the other hand, if the mutual impedance is such as to reduce the current the gain will be less than if the antennas were not coupled.

The calculation of mutual impedance between antennas is a difficult problem, and data are available only for a few special cases. Two simple, but important, ones are shown in Figs. 4-7 and 4-8. These graphs do not show the mutual impedance but instead show a more useful quantity, the radiation resistance measured at the center of an antenna as it is affected by the spacing between two antennas.

As shown by the solid curve in Fig. 4-7, the radiation resistance at the center of either dipole, when the two are self-resonant, parallel, and operated in phase, decreases rapidly as the spacing between them is increased until the spacing is about 0.7 wavelength. The maximum gain is secured from a pair of such elements when the spacing is in this region, because the current is larger for the same power and the fields from the two arrive in phase at a distant point placed on a line perpendicular to the line joining the two antennas (see Fig. 2-19, Chapter Two).

The broken curve in this figure, representing two antennas operated 180 degrees out of phase (end-fire), cannot be interpreted quite so simply. The radiation resistance decreases with decreasing spacing in this case. However, the fields from the two antennas add up in phase at a distant point in the favored direction only when the spacing is one-half wavelength (in the range of spacings considered). At smaller spacings the fields become increasingly out of phase, so the total field is less than the simple sum of the two. The latter factor decreases the gain at the same time that the reduction in radiation resistance is increasing it. As shown later in this chapter, the gain goes through a maximum when the spacing is in the region of ⅛ wavelength.

The curve for two collinear elements in phase, Fig. 4-8, shows that the radiation resistance decreases and goes through a broad minimum in the region of 0.3- to 0.5-wavelength spacing between the adjacent ends of the antennas. Since the minimum is not significantly less than the radiation resistance of an isolated antenna, the gain will not exceed the gain calculated on the basis of uncoupled antennas. That is, the best that two collinear elements will give, even with the optimum spacing, is a power gain of about 2 (3 db.). When the separation between the ends is very small—the usual method of operation—the gain is reduced.

Driven Arrays

Driven arrays may be either broadside or end-fire, and may consist of collinear elements, parallel elements, or a combination of both. The number of elements that it is practicable to use depends on the frequency and the space available for the antenna. Fairly elaborate arrays, using as many as 16 or even 32 elements, can be installed in a rather small space when the operating frequency is in the v.h.f. range. At lower frequencies the construction of antennas with a large number of elements would be impracticable for most amateurs.

It is characteristic of broadside arrays that the power gain is proportional to the length of the array but is substantially independent of the number of elements used, provided the optimum element spacing is not exceeded. This means, for example, that a 5-element array and a 6-element array will have the same gain, provided the elements in both are spaced so that the over-all array length is the same. Although this principle is seldom used for the purpose of reducing the number of elements, because of complications introduced in feeding power to each element in the proper phase, it does illustrate the fact that there is nothing to be gained by increasing the number of elements if the space occupied by the antenna is not increased proportionally.

Generally speaking, the maximum gain in the smallest linear dimensions will result when the antenna combines both broadside and end-fire directivity and uses both parallel and collinear elements. In this way the antenna is spread over a greater volume of space, which has the same

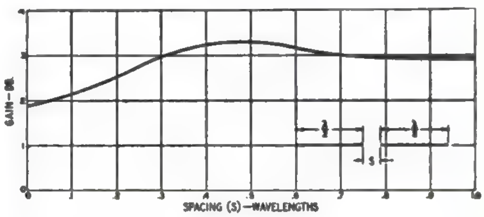


Fig. 4-9—Gain of two collinear half-wave elements as a function of spacing between the adjacent ends.

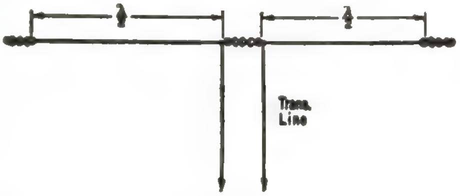


Fig. 4-10—A two-element collinear array ("two half waves in phase"). The transmission line shown would operate as a tuned line. A matching section can be substituted and a nonresonant line used if desired.

effect as extending its length to a much greater extent in one linear direction.

Feeding Driven Arrays

Not the least of the problems encountered in constructing multielement driven arrays is that of supplying the required amount of power to each element and making sure that the currents in the elements are in the proper phase. The directive patterns given in this chapter are based on the assumption that each element carries the same current and that the phasing is exact. If the element currents differ, or if the phasing is not proper, the actual directive patterns will not be quite like those shown. Small departures will not greatly affect the gain, but may increase the beam width and introduce minor lobes—or emphasize those that exist already.

If the directive properties of beam antennas are to be fully realized, care must be used to prevent antenna currents from flowing on transmission lines (see Chapter Three) used as interconnections between elements, as well as on the main transmission line. If radiation takes place from these lines, or if signals can be picked up on them, the directive effects may be masked by such stray radiation or pick-up. Although this may not greatly affect the gain either in transmission or reception, received signals coming from undesired directions will not be suppressed to the extent that is possible with a well-designed system.

COLLINEAR ARRAYS

Collinear arrays are always operated with the elements in phase. (If alternate elements in such an array are out of phase, the system simply becomes a harmonic-type antenna.) A collinear array is a broadside radiator, the direction of maximum radiation being at right angles to the line of the antenna.

Power Gain

Because of the nature of the mutual impedance between collinear elements the radiation resistance is increased as shown in Fig. 4-8. For this reason the power gain does not increase in direct proportion to the number of elements. The gain with two elements as the spacing between them is varied is shown by Fig. 4-9. Although the gain is greatest when the end-to-end spacing is in the region of 0.3 to 0.5 wavelength, the use of spacings of this order is inconvenient constructionally and introduces problems in feeding the two elements. As a result, collinear elements are almost always operated with their ends quite close together—in wire antennas, usually with just a strain insulator between.

With very small spacing between the ends of adjacent elements the theoretical power gain of collinear arrays is approximately as follows:

2 collinear elements—1.9 db.
3 collinear elements—3.2 db.
4 collinear elements—4.3 db.

More than four elements are rarely used.

Directivity

The directivity of a collinear array, in a plane containing the axis of the array, increases with its length. Small secondary lobes appear in the pattern when more than two elements are used, but the amplitudes of these lobes are low enough so that they are not important. In a plane at right angles to the array the directive diagram is a circle, no matter what the number of elements. Collinear operation, therefore, affects only the directivity in the plane containing the antenna. At right angles to the wire the pattern is the same as that of the half-wave elements of which it is composed.

When a collinear array is mounted with the elements vertical the antenna radiates equally well in all geographical directions. An array of such "stacked" collinear elements tends to confine the radiation to low vertical angles. For purposes of estimating the effect of ground reflection the height is taken as the height of the center of the array. Applying the ground-reflection factor for this height (using the reflection factors given in Chapter Two for half-wave antennas)

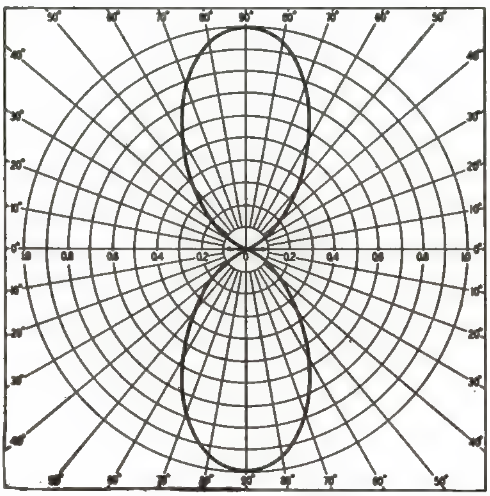


Fig. 4-11—Free-space directive diagram for a two-element collinear array. Field strength is shown on a relative basis. This is the horizontal pattern at low wave angles when the array is horizontal.

to the directive pattern of the array will give the resultant vertical pattern, taking into account ground reflection.

If a collinear array is mounted horizontally, the directive pattern in the vertical plane at right angles to the array is the same as the vertical pattern of a simple half-wave antenna at the same height (Chapter Two).

Two-Element Array

The simplest and most popular collinear array is one using two elements, as shown in Fig. 4-10. This system is commonly known as "two half waves in phase," and the manner in which the desired current distribution is secured has been described in Chapter Three. The directive pattern in a plane containing the wire axis is shown in Fig. 4-11.

Depending on the conductor size, height, and similar factors, the impedance at the feed point can be expeected to be in the range from about 4000 to 6000 ohms, for wire antennas. If the elements are made of tubing having a low length/diameter ratio, values as low as 1000 ohms are representative. The system can be fed through an open-wire tuned line with negligible loss for ordinary line lengths, or a matching section may be used if desired.

Three- and Four-Element Arrays

When more than two collinear elements are used it is necessary to connect "phasing" stubs between adjacent elements in order to bring the currents in all elements in phase. It will be recalled from Chapter Two that in a long wire the direction of current flow reverses in each half-wave section. Consequently, collinear elements cannot simply be connected end to end; there must be some means for making the current flow in the same direction in all elements. In Fig. 4-12A the direction of current flow is correct in the two left-hand elements because the transmission line is connected between them. The phasing stub between the second and third elements makes the instantaneous current direction correct in the third element. This stub may be looked upon simply as the alternate half-wave section of a long-wire antenna folded back on itself to cancel its radiation. In Fig. 4-12A the part to the right of the transmission line has a total length of three half wavelengths, the center half wave being folded back to form a quarter-wave phase-reversing stub. No data are available on the impedance at the feed point in this arrangement, but various considerations indicate that it should be over 1000 ohms.

An alternative method of feeding three collinear elements is shown in Fig. 4-12B. In this case power is applied at the center of the middle elment and phase-reversing stubs are used between this element and both of the outer elements. The impedance at the feed point in this case is somewhat over 300 ohms and provides a close match to 300-ohm line. The s.w.r. will

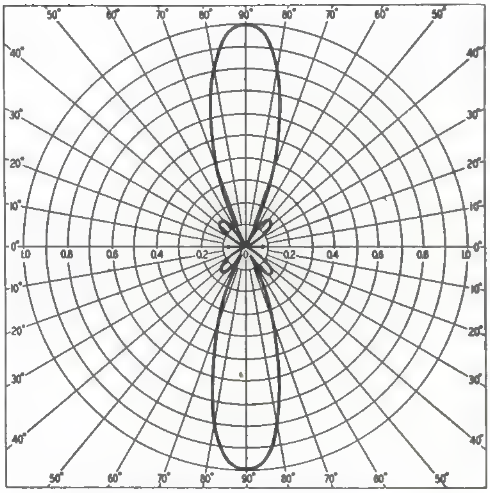


Fig. 4-13—Free-space directive diagram for a four-element collinear array. Field strength is shown on a relative basis.

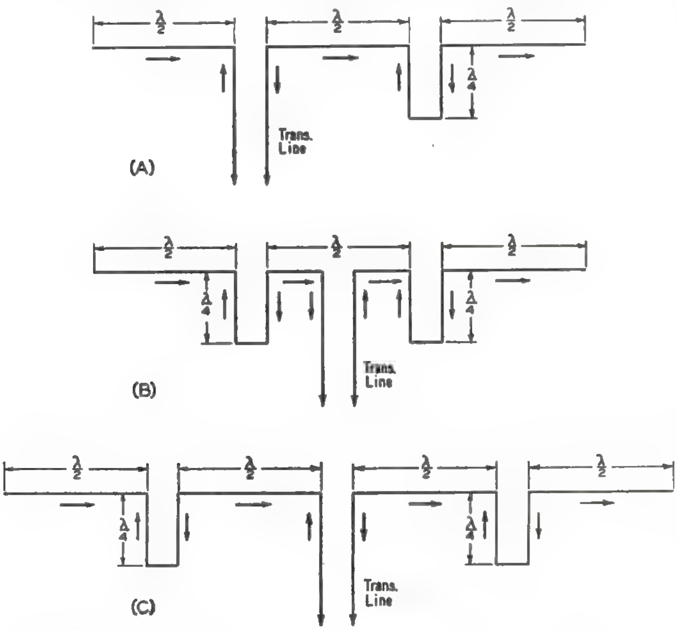


Fig. 4-12—Three- and four-element collinear arrays. Alternative methods of feeding a three-element array are shown at A and B. These drawings also show the current distribution on the antenna elements and phasing stubs. A matched transmission line can be substituted for the tuned line by using a suitable matching section.

be less than 2 to 1 when 600-ohm line is used. Center feed of this type is somewhat preferable to the arrangement in Fig. 4-12A because the system as a whole is balanced. This assures more uniform power distribution among the elements. In A, the right-hand element is likely to receive somewhat less power than the other two because a portion of the fed power is radiated by the middle element before it can reach the one located at the extreme right.

A four-element array is shown in Fig. 4-12G. The system is symmetrical when fed between the two center elements as shown. As in the three-element case, no data are available on the impedance at the feed point. However, the s.w.r. with a 600-ohm line should not be much over 2 to 1. Fig. 4-13 shows the directive pattern of a four-element array. The sharpness of the three-element pattern is intermediate between Figs. 4-11 and 4-13, with a small minor lobe at right angles to the array axis.

Collinear arrays can be extended to more than four elements. However, the simple two-element

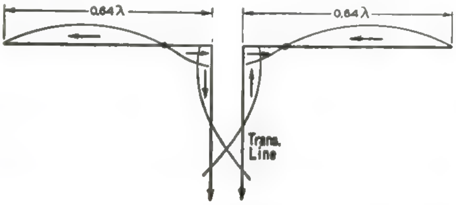


Fig. 4-14—The extended double Zepp. This system gives somewhat more gain than two half-wave collinear elements.

collinear array is the type most used, for the reason that it lends itself well to multiband operation. More than two collinear elements are seldom used because more gain can be obtained from other types of arrays.

Adjustment

In any of the collinear systems described the lengths of the radiating elements can be found from the formula $468/f$ (MC). The lengths of the phasing stubs can be found from the formulas given in Chapter Three for the type of line used. If the stub is open-wire line (500 to 600 ohms impedance) it is satisfactory to use a velocity factor of 0.975 in the formula for a quarter-wave line. On-the-ground adjustment is, in general, an unnecessary refinement. If desired, however, the following procedure may be used when the system has more than two elements:

Disconnect all stubs and all elements except those directly connected to the transmission line (in the case of feed such as is shown in Fig. 4-12B leave only the center element connected to the line). Adjust the elements to resonance, preferably using the still-connected element or method described in Chapter Three. When the proper length is determined, cut all other elements to the same length. Make the phasing stubs slightly long and use a shorting bar to

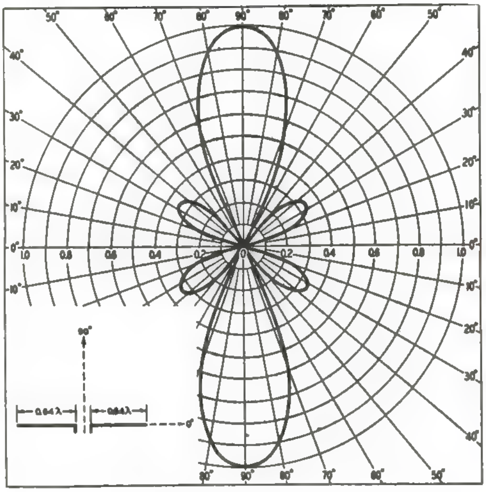


Fig. 4-15—Free-space directive diagram for the extended double Zepp. This is also the horizontal directional pattern when the elements are horizontal.

adjust their length. Connect the elements to the stubs and adjust the stubs to resonance, as indicated by maximum current in the shorting bars or by the positions of the standing waves along the transmission line. When the whole system is resonant the position of the first current or voltage maximum along the transmission line should be the same as when the line is shorted or open, as described in Chapter Three. If more than three or four elements are used it is best to add elements two at a time (one at each end of the array), resonating the system each time before a new pair is added.

The Extended Double Zepp

An expedient that may be adopted to obtain the higher gain that goes with wider spacing in a simple system of two collinear elements is to make the elements somewhat longer than ½ wavelength. As shown in Fig. 4-14, this increases the spacing between the two in-phase half-wave sections at the ends of the wires. The section in the center carries a current of opposite phase, but if this section is short the current will be small because it represents only the outer ends of a half-wave section. Because of the small current and short length the radiation from the center is small. The optimum length for each element is 0.64 wavelength. At greater lengths the system tends to act as a long-wire antenna and the gain decreases.

This system is known as the "extended double Zepp." The gain over a half-wave dipole is approximately 3 db., as compared with slightly less than 2 db. for two collinear dipoles. The directional pattern in the plane containing the axis of the antenna is shown in Fig. 4-15. As in the case of all other collinear arrays, the free-space pattern in the plane at right angles to the antenna elements is the same as that of a half-wave antenna; i.e., is circular.

BROADSIDE ARRAYS WITH PARALLEL ELEMENTS

To obtain broadside directivity with parallel elements the currents in the elements must all be in phase. At a distant point lying on a line perpendicular to the axis of the array and also perpendicular to the plane containing the elements, the fields from all elements add up in

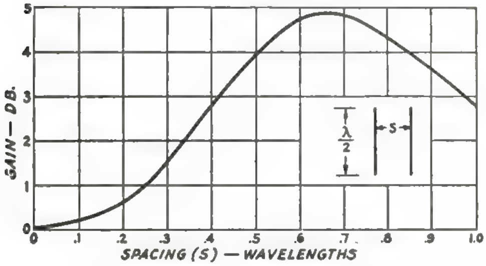


Fig. 4-16—Power gain as a function of the spacing between two parallel elements operated in phase (broadside).

phase. The situation is similar to that pictured in Fig. 4-1 in this chapter and in Fig. 2-19, Chapter Two.

Broadside arrays of this type theoretically can have any number of elements. However, practical limitations of construction and available space usually limit the number of broadside parallel elements to two, in the amateur bands below 30 Mc., when horizontal polarization is used. More than four such elements seldom are used even at v.h.f.

Power Gain

The power gain of a parallel-element broadside array depends on the spacing between elements as well as on the number of elements. The way in which the gain of a two-element array varies with spacing is shown in Fig. 4-16. The greatest gain is obtained when the spacing is in the vicinity of 0.7 wavelength.

The theoretical gains of broadside arrays having more than two elements are approximately as follows:

No. of Parallel Elements	*Db. Gain with ½-Wave Spacing*	*Db. Gain with ¾-Wave Spacing*
3	5	7
4	6	8.5
5	7	10
6	8	11

The element must, of course, all lie in the same plane.

Directivity

The sharpness of the directive pattern depends on the spacing between elements and on the number of elements. Larger element spacing will sharpen the main lobe, for a given number of elements. The two-element array has no minor lobes when the spacing is ½ wavelength, but small minor lobes appear at greater spacings. When three or more elements are used the pattern always has minor lobes.

The vertical directive pattern of such an array when the elements are vertical is the same as that for a simple half-wave dipole at the same height. The patterns are given in Chapter Two. When the array elements are horizontal the vertical pattern is the product of the broadside pattern for the particular array used multiplied by the ground-reflection factors given in Chapter Two. For the purpose of applying the ground-reflection factor the height of the array is taken as the mean height above ground. The horizontal directive pattern of a horizontally-polarized parallel-element broadside array is the same as that of a simple dipole.

Two-Element Arrays

The elements of a broadside array must be connected by transmission lines that supply power in the proper phase to each element. Three methods of interconnection for a two-element array are given in Fig. 4-17. In A, the main transmission line is connected to the "phasing line" at its center. The two halves of the phasing line, *AB* and *AC*, are simply in parallel, insofar as the main tranmission line is concerned, so the currents in the phasing line flow in opposite directions, with respect to the junction *A*.

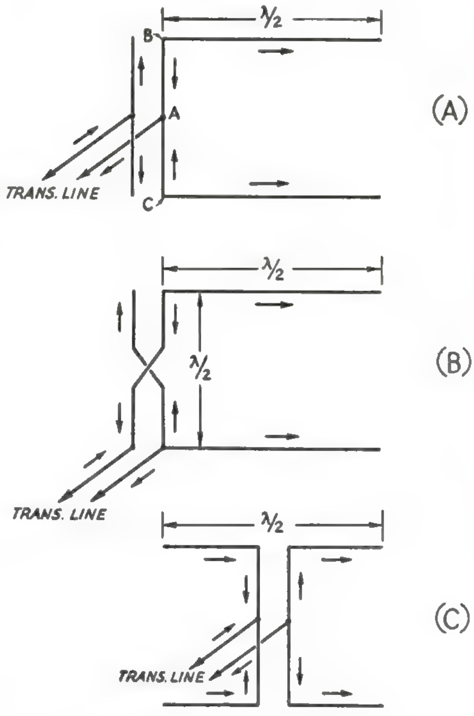


Fig. 4-17—Two-element broadside arrays, showing different methods of supplying power.

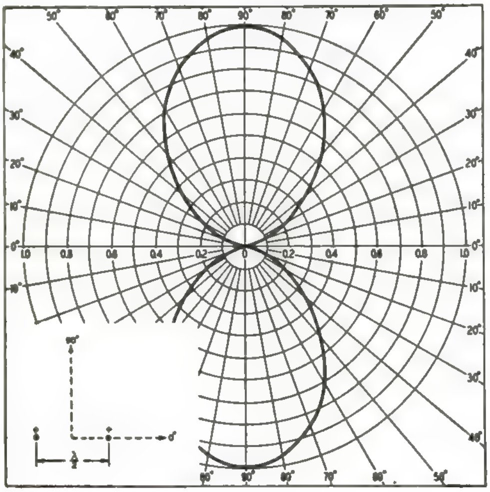


Fig. 4-18—Free-space directive diagram of a two-element broadside array for an element spacing of ½ wavelength. This drawing gives the low-angle horizontal pattern of a vertically-polarized array.

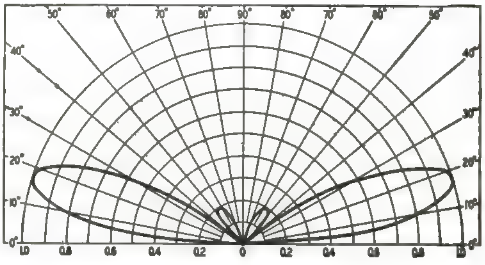


Fig. 4-19—Vertical pattern broadside to a two-element in-phase array with horizontal elements. This pattern is for a mean height of ¾ wavelength; i.e., lower element ½ wavelength high and upper element one wavelength high. At low wave angles, the horizontal pattern of such an array is the same as for a half-wave dipole.

This brings the currents in the array elements in phase. The phasing line can be any convenient length in this case, so the line spacing between the two elements can be any value desired. Although no data are available on impedances, a rough estimate indicates that in most practical cases the impedance will be well below 100 ohms at the point where the main transmission line joins the phasing line, assuming a half-wave phasing line having a Z_0 of about 600 ohms. If the phasing line is not exactly ½ wavelength long the impedance will be reactive as well as resistive.

In B, the main transmission line is connected at the junction of the phasing line and one element. In this case it is necessary to transpose the phasing line somewhere along its length so that the element currents will be in the proper phase. This is shown by the arrows indicating relative direction of current flow. The impedance at the feed point will be resistive and of the order of a few thousand ohms when the elements and phasing line all have electrical lengths of ½ wavelength. With this type of feed the spacing between the elements is determined by the electrical length of the phasing line; it must be an *electrical* half wave long to bring the element currents in proper phase. Open-wire lines are always used as phasing lines in this type of system because their electrical length is nearest to the length of an actual half wavelength in space. If the velocity factor of the phasing line is much less than 1 the antenna elements will perforce be considerably less than a half wavelength apart, and this will reduce the gain.

A third method of feeding is shown at C. This is the best of the three, insofar as symmetry is concerned. The spacing between the two elements can be any desired value. However, when the spacing is one-half wavelength the impedance at the point where the main transmission line is connected is resistive and can be calculated with the aid of Fig. 4-7. For example, Fig. 4-7 shows that the radiation resistance of each element is approximately 60 ohms at half-wave spacing. If the Z_0 of the phasing line is 600 ohms, the impedance reflected at the transmission-line terminals will be $(600)^2/60 = 6000$ ohms, since with half-wave spacing the phasing line is ¼ wave long from the element to the junction. As the reflected resistances from both elements are in parallel, the resistive impedance seen by the transmission line is $600/2 = 3000$ ohms.

The arrays shown in Fig. 4-17 may be installed either vertically or horizontally, depending on the type of polarization desired. The free-space directive diagram given in Fig. 4-18 is also the horizontal pattern of the array at low wave angles when the elements are vertical. The vertical pattern for a horizontally-polarized two-element array for a mean height of ¾ wavelength is given in Fig. 4-19. The pattern for other heights may be found by multiplying the pattern of Fig. 4-18 by the ground-reflection factor for the actual mean height.

Multielement Arrays

Three- and 4-element arrays are shown in Fig. 4-20. In the 3-element array with half-wave spacing (A) the array is fed at the center. This is the most desirable point in that it tends to keep the power distribution between elements uniform. However, the transmission line could be connected at either *B* or *C*.

When the spacing is greater than ½ wavelength the phasing lines must be one wavelength long and are not transposed between elements. This is shown at B in Fig. 4-20. With this arrangement any element spacing up to one wavelength can be used, if the phasing lines can be folded as suggested in the drawing.

The 4-element array at C is fed at the center of the system to make the power distribution between elements as uniform as possible. However, the transmission line could be connected at either *B*, *C*, *D* or *E*. In such case the section of phasing line between *B* and *D* must be transposed in order to make the currents flow in the same direction in all elements. The 4-element array at C and the 3-element array at B have approximately the same gain when the element spacing in the latter is ¾ wavelength.

An alternative feeding method is shown at D. This system can also be applied to the 3-element arrays, and will result in better symmetry in any case. It is only necessary to move the phasing line to the center of each element, making connection to both sides of the line instead of one only.

The free-space pattern for a 4-element array with half-wave spacing is shown in Fig. 4-21. This is also approximately the pattern for a 3-element array with ¾-wave spacing. The major lobe of a 3-element array with half-wave spacing is intermediate in sharpness between a 2-element and a 4-element array.

Larger arrays can be designed and constructed by following the phasing principles shown in the drawings.

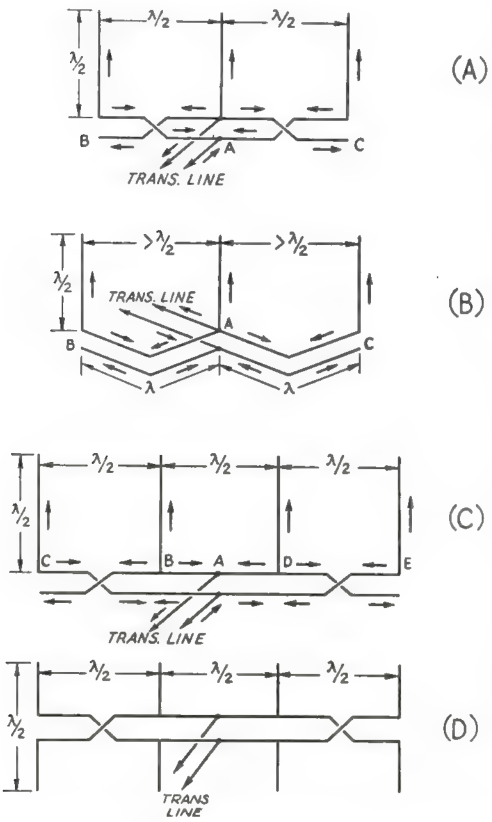


Fig. 4-20—Methods of feeding three- and four-element broadside arrays with parallel elements.

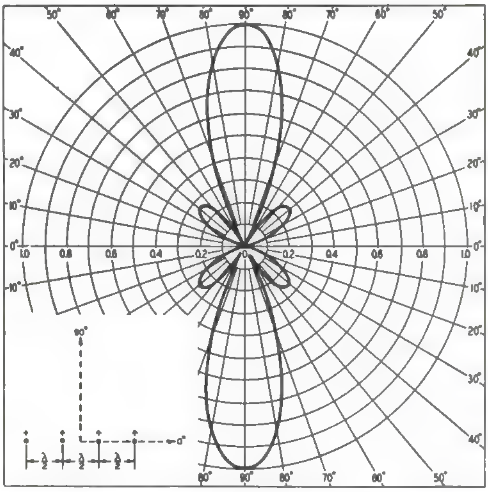


Fig. 4-21—Free-space directive diagram of a four-element broadside array using parallel elements. This is also the horizontal directive pattern at low wave angles for a vertically-polarized array.

No accurate figures are available for the impedances at the various feed points indicated in the drawings. It can be estimated to be in the vicinity of 1000 ohms when the feed point is at a junction between the phasing line and a half-wave element, becoming smaller as the number of elements in the array is increased. When the feed point is midway between end-fed elements as in Fig. 4-20C, the impedance of a 4-element array as seen by the transmission line is in the vicinity of 200-300 ohms, with 600-ohm open-wire phasing lines. The impedance at the feed point with the antenna shown at D should be about 1500 ohms.

END-FIRE ARRAYS

The term "end-fire" covers a number of different methods of operation, all having in common the fact that the maximum radiation takes place along the array axis, and that the array consists of a number of parallel elements in one plane. End-fire arrays can be either bidirectional or unidirectional. In the bidirectional type commonly used by amateurs there are only two elements, and these are operated with currents 180 degrees out of phase. Unidirectional end-fire driven arrays have not had much amateur use because the element phasing is neither 0 nor 180 degrees and tends to be complicated from an adjustment standpoint. (Instead, unidirectional antennas as used by amateurs are practically all based on the use of parasitic elements as described later in this chapter.)

Two-Element Arrays

In the two-element array with equal currents out of phase the gain varies with the spacing between elements as shown in Fig. 4-22. The maximum gain is in the neighborhood of ⅛-wave

spacing. Below 0.05-wave spacing the gain decreases rapidly, since the system is approaching the spacings used for nonradiating transmission lines.

The radiation resistance at the center of either element is very low at the spacings giving the greatest gain, as shown by Fig. 4-7. The spacings most frequently used are ⅛ and ¼ wavelength, at which the resistances are about 8 and 32 ohms, respectively. When the spacing is ⅛ wavelength

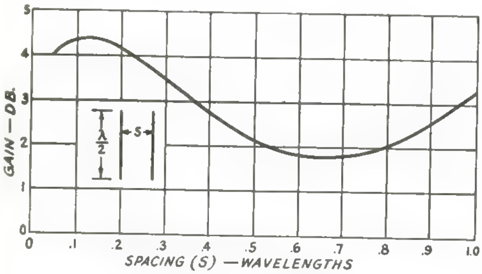


Fig. 4-22—Gain of an end-fire array consisting of two elements fed 180 degrees out of phase, as a function of the spacing between elements. Maximum radiation is in the plane of the elements and at right angles to them at spacings up to 0.5 wavelength, but the direction changes at greater spacings.

it is advisable to use good-sized conductors—preferably tubing—for the elements because with the radiation resistance so low the heat losses in the conductor can represent an appreciable portion of the power supplied to the antenna. Excessive conductor loss will mean that the theoretical gain cannot be realized.

Three methods of feeding bidirectional end-fire elements are shown in Fig. 4-23. In A, one section of the phasing line is transposed to bring the element currents in proper phase. The method at B is suitable for close-spaced (i.e., ⅛-wave) arrays because each half of the connecting wire is only 1/16 wave long and carries very little current. Hence there is very little radiation from the wires joining the ends of the elements to the transmission line even though the currents are in phase. The center-fed arrangement shown at C is especially useful when the antenna is to be operated on two bands—for example, 14 and 28 Mc.—the higher of which is the second harmonic of the lower.

Because of the very low radiation resistance when the spacing, S in Fig. 4-23, is ⅛ wavelength, the s.w.r. on the transmission line is very high. No figures are available for the end-fed cases, but it can be estimated to be 20 to 1 or higher. With center feed using a 600-ohm line the s.w.r. is over 30 to 1. If the transmission line has any considerable distance to run, it is advisable to match it to the antenna by using a matching section of the type described in Chapter Three. Such a matching section should be of open-wire construction in view of the high s.w.r. The line itself, of course, can be any type capable of carrying the transmitter power. If the transmission line does not have to run more than a wavelength or two it may be of open-wire construction and operated as a tuned line.

With ¼-wave spacing the increased radiation resistance will lower the s.w.r. considerably. With center feed it will be about 10 to 1 (600-ohm line), and should not exceed that figure with end feed.

In a close-spaced array fed through a tuned transmission line the element lengths are not critical; the only point to watch is to preserve the symmetry of the system as a whole. When a matching section is used, however, it is necessary to adjust the system accurately to the particular frequency to be used most. The low radiation resistance makes the antenna a sharply-tuned affair, and so relatively small departures from the design frequency will throw off the impedance match.

Another way of overcoming the high s.w.r. on the transmission line, and at the same time reducing the resistance loss in the antenna elements, is to use a folded-dipole (see Chapter Three) arrangement as indicated at D in Fig. 4-23. In this way the impedance at the element

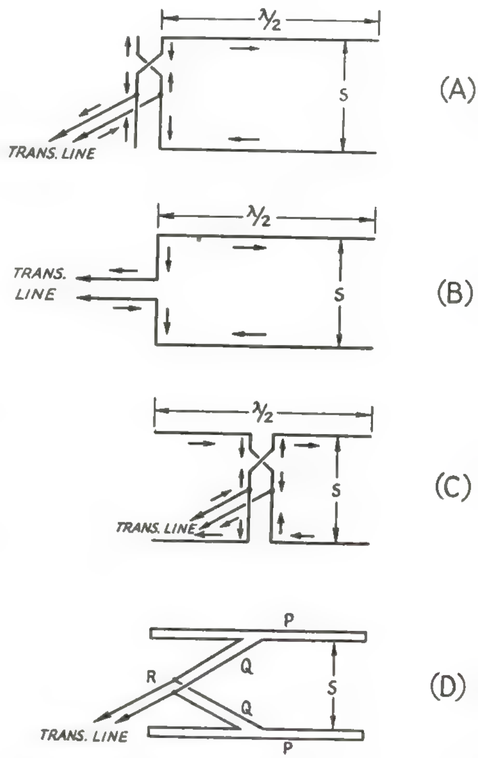


Fig. 4-23—Parallel-element end-fire array with various methods of feed, showing current distribution. Matching sections for making the transmission line nonresonant, as described in Chapter Three, may be used in the first three cases. The distance S may be selected from Fig. 4-22.

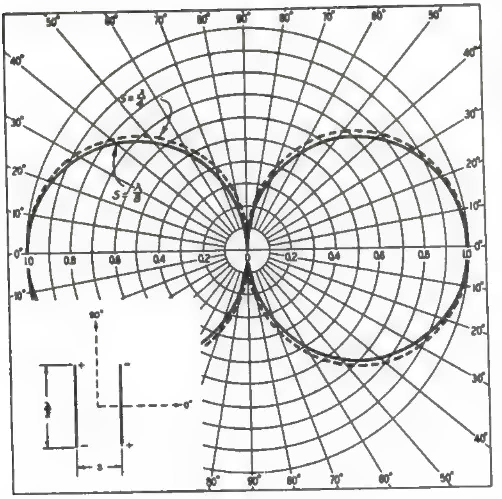


Fig. 4-24—Free-space directive diagram of a two-element end-fire array with 180-degree phasing, in the plane containing the two parallel elements.

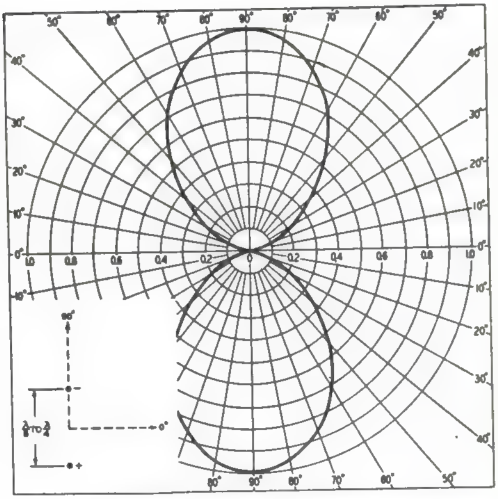


Fig. 4-25—Free-space directive diagram of a two-element end-fire array with 180-degree phasing, in the plane at right angles to the plane containing the elements.

terminals is stepped up, and then there is a further impedance step-up in the section Q, which is a quarter-wave Q-type matching transformer. A number of combinations are listed below:
In each case the s.w.r. on the transmission line will be well below 2 to 1. Twin-Lead will be satisfactory for the 75- and 300-ohm line. The velocity factor of the line must be taken into account in determining the physical length of the ¼-wave matching transformer. In all the ar-

S, wave-length	*No. of conductors in dipole*	*Z_0 of ¼-wave matching section*	*Z_0 of main transmission line*
⅛	1	75	300
⅛	2	75	75
⅛	3	300	600
⅛	4	300	300
¼	1	75	75
¼	2	300	300
¼	3	600	600

rangements listed above except those using plain dipoles as elements the frequency characteristic of the antenna will be broadened somewhat by the folded-dipole construction.

The free-space directive pattern in the plane containing the array is given in Fig. 4-24 and the corresponding pattern in the plane at right angles to the array plane is given in Fig. 4-25 Fig. 4-24 is also the horizontal directive pattern of the array at low wave angles when the elements are horizontal, while Fig. 4-25 is the horizontal pattern at low wave angles when the elements are vertical. The vertical pattern of a horizontally-polarized array is shown in Fig. 4-26.

Unidirectional End-Fire Arrays

Two parallel elements spaced ¼ wavelength apart and fed equal currents 90 degrees out of

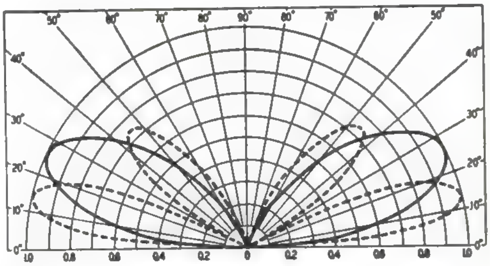


Fig. 4-26—Vertical pattern of a horizontally-polarized two-element end-fire array. Solid curve, height ½ wavelength; broken curve, height 1 wavelength.

phase will have a directional pattern, in the plane at right angles to the plane of the array, as represented in Fig. 4-27. The maximum radiation is in the direction from the element in which the current leads to the element in which the current lags. In the opposite direction the fields from the two elements cancel.

One way in which the 90-degree phase difference can be obtained is shown in Fig. 4-28. Each element must be matched to its transmission line, the two lines being of the same type except that one is an electrical quarter wavelength longer than the other. The length L can be any convenient value. Open quarter-wave matching sections are shown, but half-wave shorted sections could be used instead. The two transmission lines are connected in parallel at the transmitter coupling circuit.

When the currents in the elements are neither in phase nor 180 degrees out of phase the radiation resistances of the elements are not equal. This complicates the problem of feeding equal currents to the elements. If the currents are not

equal one or more minor lobes will appear in the pattern and decrease the front-to-back ratio. The adjustment process is likely to be tedious and requires field-strength measurements in order to get the best performance.

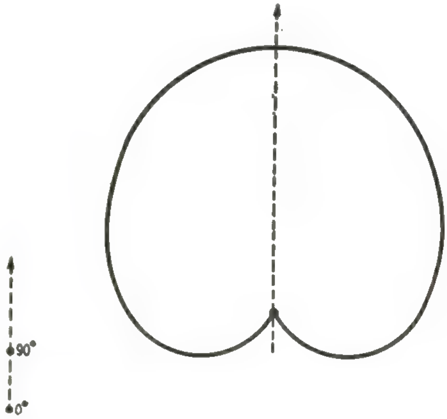


Fig. 4-27—Representative pattern for a two-element end-fire array with 90-degree phasing, in the plane perpendicular to the plane containing the elements.

More than two elements can be used in a unidirectional end-fire array. The requirement for unidirectivity is that there must be a progressive phase shift in the element currents equal to the spacing, in electrical degrees, between the elements, and the amplitudes of the currents in the various elements also must be properly related. This requires "binomial" current distribution—i.e., the ratios of the currents in the elements must be proportional to the coefficients of the binomial series. In the case of three elements, a type of antenna that has been used in amateur work (Andrew, *QST*, July, 1950), this requires that the current in the center element be twice that in the two outside elements, for 90-degree

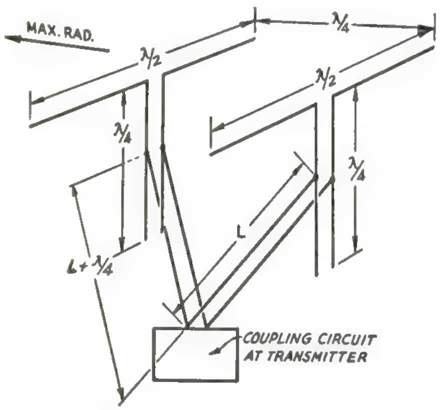


Fig. 4-28—Unidirectional two-element end-fire array and method of obtaining 90-degree phasing.

(quarter-wave) spacing and element-current phasing. This antenna has an overall length of ½ wavelength which, although rather large for 14 Mc., has been used in rotatable antennas for that band. The constructional difficulties are of course much less acute at higher frequencies.

COMBINATION DRIVEN ARRAYS

Broadside, end-fire and collinear elements can readily be combined to increase gain and directivity, and this is in fact usually done when more than two elements are used in an array. Combinations of this type give more gain, in a given amount of space, than plain arrays of the types just described.

The combinations that can be worked out are almost endless, but in this section we shall describe only a few of the simpler types that are in common use. The drawings that follow all show the elements arranged for horizontal polarization, which is customary on the frequencies below 30 Mc. where these arrays find their greatest application. For vertical polarization the arrays should be rotated 90 degrees so that the elements are vertical—that is, "stood on end."

Other methods of interconnecting elements than those shown in the drawings may be used. However, the methods shown are recommended over others for two reasons: The antenna system is symmetrical with respect to the feed point, thus making the current distribution among elements as uniform as possible; the lengths of phasing lines (and antenna elements as well) are not critical so long as the lengths of lines radiating from a junction are all the same. With other feed methods this may not be true, and it becomes necessary to use the methods described in Chapter Three to ensure that elements and phasing lines are exactly resonant at the design frequency, if maximum performance is to be secured from the antenna. This adjustment process can be rather difficult as well as tedious. If the feed arrangements shown in the drawings are followed the lengths of wire elements can be found from $468/f$ (Mc.), the element spacings from $984/f$ (Mc.) multiplied by the fraction of wavelength desired, and the phasing lines can simply be cut to fit, keeping all lines the same length.

Gain of Combination Arrays

The accurate calculation of the power gain of a multielement array requires a knowledge of the mutual impedances between all elements. For approximate purposes it is sufficient to assume that each *set* (collinear, broadside, end-fire) will have the gains as given earlier, and then simply add up the gains for the combination. This neglects the effects of cross-coupling between sets of elements. However, the array configurations are such that the mutual impedances from cross-coupling should be relatively small, particularly when the spacings are ¼ wavelength or more, so that the estimated gain should be reasonably close to the actual gain.

Four-Element End-Fire and Collinear Array

The array shown in Fig. 4-29 combines collinear in-phase elements with parallel out-of-phase elements to give both broadside and end-fire directivity. It is popularly known as a "two-section W8JK" or "two-section flat-top beam." The approximate gain calculated as described above is 6.2 db. with ⅛-wave spacing and 5.7 db. with ¼-wave spacing. Directive patterns are given in Figs. 4-30 and 4-31.

The impedance between elements at the point where the phasing line is connected is of the order of several thousand ohms. The s.w.r.

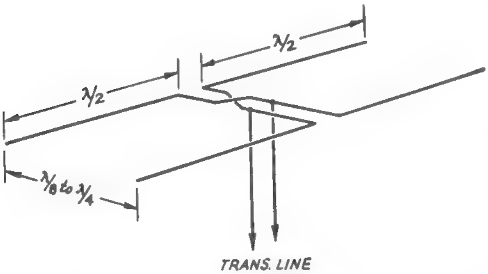


Fig. 4-29—A four-element array combining collinear broadside elements and parallel end-fire elements.

with an unmatched line consequently is quite high, and this system should be constructed with open-wire line (500 or 600 ohms) if the line is to be resonant. To use a matched line a closed stub 3/16 wavelength long can be connected at the transmission-line junction shown in Fig. 4-29, and the transmission line itself can then be tapped on this matching section at the point resulting in the lowest line s.w.r. This point can be determined by trial.

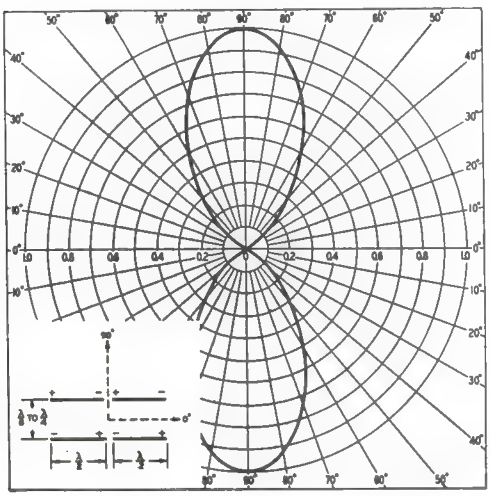


Fig. 4-30—Free-space directive diagram of the antenna shown in Fig. 4-29, in the plane of the antenna elements. The pattern in the plane perpendicular to the element plane is the same as Fig. 4-25.

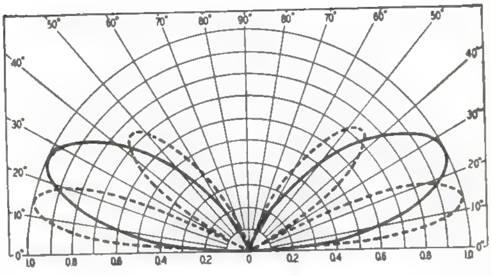


Fig. 4-31—Vertical pattern of the four-element antenna of Fig. 4-29 when mounted horizontally. Solid curve, height ½ wavelength; broken curve, height 1 wavelength. Fig. 4-30 gives the horizontal pattern.

With ¼-wave spacing the s.w.r. on a 600-ohm line is estimated to be in the vicinity of 3 or 4 to 1.

This type of antenna can be operated on two bands having a frequency ratio of 2 to 1, if a resonant feed line is used. For example, if designed for 28 Mc. with ¼-wave spacing between elements it can be operated on 14 Mc. as a simple end-fire array (Fig. 4-23C) having ⅛-wave spacing.

Four-Element Broadside Array

The four-element array shown in Fig. 4-32 is commonly known as the "lazy-H." It consists of a set of two collinear elements and a set of two parallel elements, all operated in phase to give broadside directivity. The gain and directivity will depend on the spacing, as in the case of a simple parallel-element broadside array. The spacing may be chosen between the limits

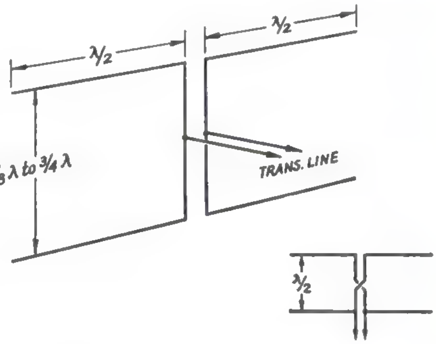


Fig. 4-32—Four-element broadside array ("lazy-H") using collinear and parallel elements.

shown on the drawing, but spacings below ⅜ wave length are not worth while because the gain is small. Estimated gains are as follows.

⅜-wave spacing — 4.4 db.
½-wave spacing — 5.9 db.
⅝-wave spacing — 6.7 db.
¾-wave spacing — 6.6 db.

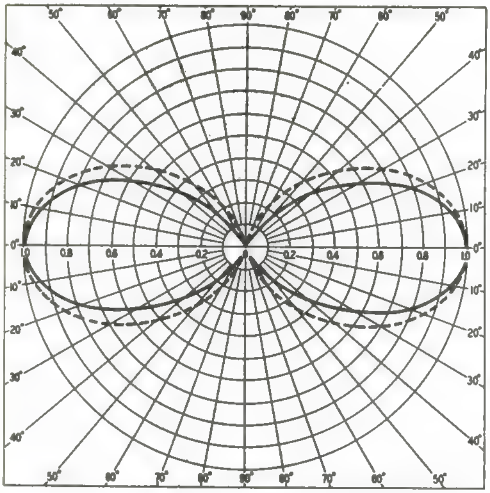


Fig. 4-33—Free-space directive diagrams of the four-element antenna shown in Fig. 4-32. The solid curve is the horizontal directive pattern at low wave angles when the antenna is mounted with the elements horizontal. The broken curve is the free-space vertical pattern of a horizontally-polarized array, broadside to the array. Actual pattern in the presence of ground may be found by multiplying this pattern by the ground-reflection factors given in Chapter Two.

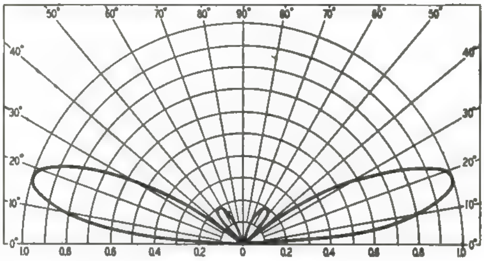


Fig. 4-34—Vertical pattern of the four-element broadside antenna of Fig. 4-32, when mounted with the elements horizontal and the lower set ½ wavelength above ground. "Stacked" arrays of this type give best results when the lowest elements are at least ½ wave high. The gain is reduced and the wave angle raised if the lowest elements are close to ground.

Half-wave spacing is generally used. Directive patterns for this spacing are given in Figs. 4-33 and 4-34.

With half-wave spacing between parallel elements the impedance at the junction of the phasing line and transmission line is resistive and is in the vicinity of 100 ohms. With larger or smaller spacing the impedance at this junction will be reactive as well as resistive. Matching stubs are recommended in cases where a nonresonant line is to be used. They may be calculated and adjusted as described in Chapter Three, after first determining the position of the current loop or node (on the transmission line) nearest the junction and after measuring the standing-wave ratio.

The system shown in Fig. 4-32 may be used on two band having a 2-to-1 frequency relationship. It should be designed for the higher of the two frequencies, using ¾-wave spacing between parallel elements. It will then operate on the lower frequency as a simple broadside array with ⅜-wave spacing.

An alternative method of feeding is shown in the small diagram in Fig. 4-32. In this case the elements and the phasing line must be adjusted exactly to an electrical half wavelength. The impedance at the feed point will be resistive and of the order of 2000 ohms.

Four-Element Broadside and End-Fire Array

The array shown in Fig. 4-35 combines parallel elements in broadside and end-fire directivity. Approximate gains can be calculated by adding the figures from Figs. 4-16 and 4-22 for the element spacings used. The smallest (physically) array—⅜-wave spacing between broadside and ⅛-wave spacing between end-fire elements—has an estimated gain of 6.8 db. and the largest—¾- and ¼-wave spacing, respectively—about 8.5 db. The optimum element spacings are ⅝ wave broadside and ⅛ wave end-fire, giving an over-all gain estimated at 9.3 db.

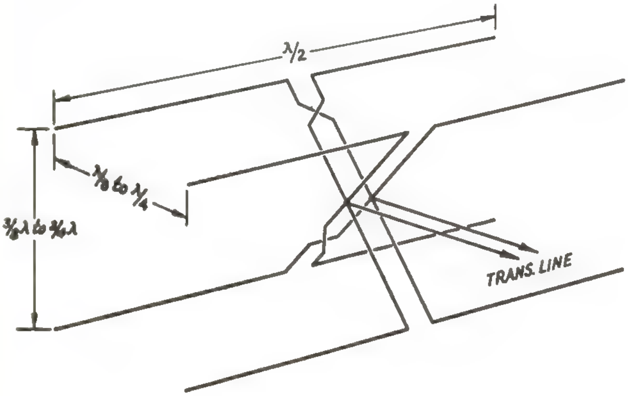


Fig. 4-35—Four-element array combining both broadside and end-fire elements.

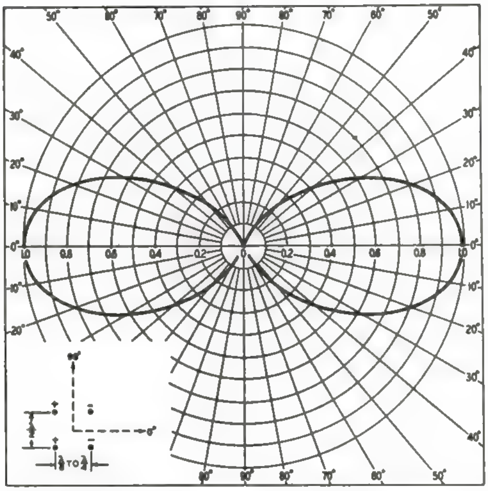


Fig. 4-36—Free-space pattern of the four-element antenna shown in Fig. 4-35, in the plane perpendicular to the array axis. The pattern in the plane containing a set of end-fire elements is the same as Fig. 4-24.

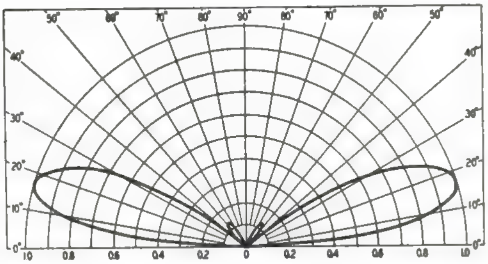


Fig. 4-37—Vertical pattern of the antenna shown in Fig. 4-35 at a mean height of ¾ wavelength (lowest elements ½ wave above ground) when the antenna is horizontally polarized. For optimum gain and low wave angle the mean height should be at least ¾ wavelength.

Directive patterns are given in Figs. 4-36 and 4-37.

The impedance at the feed point will not be purely resistive unless the element lengths are correct and the phasing lines are exactly a half wavelength long. (This requires somewhat less than half-wave spacing between broadside elements.) In this case the impedance at the junction is estimated to be over 10,000 ohms. With other element spacings the impedance at the junction will be reactive as well as resistive, but in any event the standing-wave ratio will be quite large. An open-wire line can be used as a resonant line, or a matching section may be used for nonresonant operation.

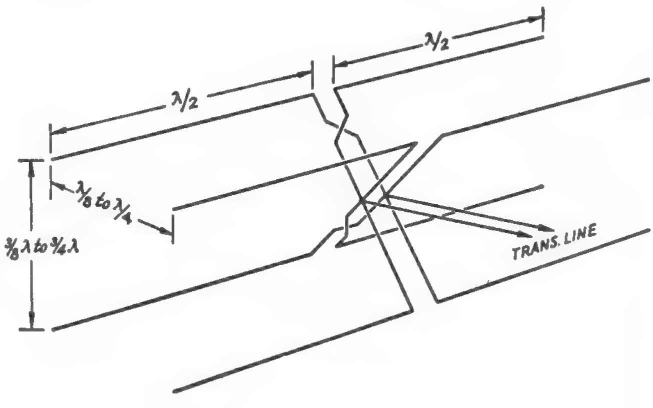
Eight-Element Driven Array

The array shown in Fig. 4-38 is a combination of collinear and parallel elements in broadside and end-fire directivity. The gain can be calculated as described earlier, usings Figs. 4-9, 4-16 and 4-22. Common practice is to use half-wave spacing for the parallel broadside elements and ⅛-wave spacing for the end-fire elements. This gives an estimated gain of about 10 db. Directive patterns for an array using these spacings are given in Figs. 4-39 and 4-40.

Although even approximate figures are not available, the s.w.r. with this arrangement will be high. Matching stubs are recommended for making the line nonresonant. Their position and length can be determined by measuring the s.w.r. and locating the current loop or null nearest the junction of the transmission and phasing lines. The procedure is described in Chapter Three.

This system can be used on two bands related in frequency by a 2-to-1 ratio, providing it is designed for the higher of the two with ¾-wave spacing between the parallel broadside elements and ¼-wave spacing between the end-fire elements. On the lower frequency it will then operate as a four-element antenna of the type shown in Fig. 4-35, with ⅜-wave broadside spacing and ⅛-wave end-fire spacing. For two-band operation a resonant transmission line will have to be used.

Fig. 4-38 — Eight-element driven array combining collinear and parallel elements for broadside and end-fire directivity.



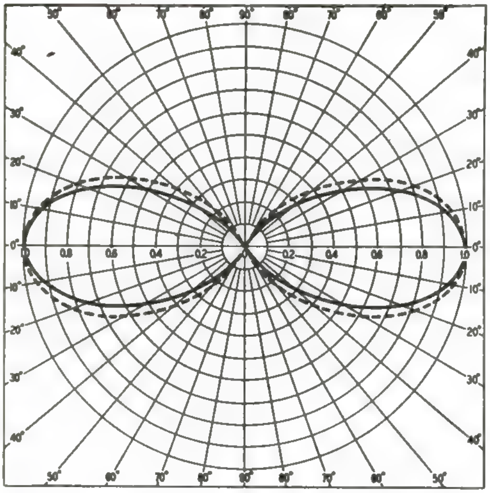


Fig. 4-39—Free-space directive diagrams of the eight-element array shown in Fig. 4-38. The solid curve is the horizontal pattern of the antenna at low wave angles when the antenna is horizontally polarized. Broken curve, free-space vertical pattern of a horizontally-polarized array, broadside to the array. The vertical pattern in the presence of ground may be found by applying the ground-reflection factors in Chapter Two, for the actual mean height.

Checking Phasing

In the antenna diagrams earlier in this chapter the relative direction of current flow in the various antenna elements and connecting lines was shown by arrows. In laying out any antenna system it is necessary to know that the phasing lines are properly connected, otherwise the antenna may have entirely different characteristics than anticipated. The phasing may be checked either on the basis of current direction or polarity of voltages. There are two rules to remember:

1) In every half-wave section of wire, starting from an open end, the current directions reverse. In terms of voltage, the polarity reverses at each half-wave point, starting from an open end.

2) Currents in transmission lines always must flow in opposite directions in adjacent wires. In terms of voltage, polarities always must be opposite.

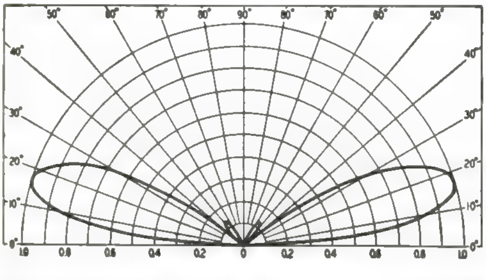


Fig. 4-40—Vertical pattern of the antenna of Fig. 4-38 when mounted horizontally at a mean height of ¾ wavelength. This is the minimum height that should be used for realizing good gain and a low wave angle.

Examples of the use of current direction and voltage polarity are given at A and B, respectively, in Fig. 4-41. The half-wave points in the system are marked by the small circles. When current in one section flows toward a circle, the current in the next section must also flow toward it, and vice versa. In the four-element antenna shown at A, the current in the upper right-hand element cannot flow toward the tranmission line, because then the current in the right-hand section of the phasing line would have to flow upward and thus would be flowing in the same direction as the current in the left-hand wire. The phasing line would simply act like two wires in parallel in such a case.

C shows the effect of transposing the phasing line. This transposition reverses the direction of current flow in the lower pair of elements, as compared with A, and thus changes the array from a combination collinear and end-fire arrangement into a collinear-broadside array.

The drawing at D shows what happens when the transmission line is connected at the center of a section of phasing line. Viewed from the main transmission line the two parts of the phasing line are simply in parallel, so the half wavelength is measured from the antenna element along the upper section of phasing line and thence along the transmission line. The distance from the lower elements is measured in the same way. Obviously the two sections of phasing line should be the same length. If they are not, the current distribution becomes quite complicated; the element currents are neither in phase nor 180 degrees out of phase, and the elements at opposite ends of the lines do not receive the same power. To change the element current phasing at D into the phasing at A, simply transpose the wires in one section of the phasing line; this reverses the direction of current flow in the antenna elements connected to that section of phasing line.

OTHER DRIVEN SYSTEMS

Two other types of driven antennas are worthy of mention, although their use by amateurs has been rather limited. The Sterba array, shown at A in Fig. 4-42, is a broadside radiator consisting of both collinear and parallel elements, with ½-wave spacing between the latter. Its distinctive feature is the method of closing the ends of the system. For direct current and low-frequency a.c. the system forms a closed loop, which is advantageous in that heating currents can be sent through the wires to melt the ice that forms in cold climates. There is comparatively little radiation from the vertical connecting wires at the ends because the currents are relatively small and are flowing in opposite directions with respect to the center (the voltage loop is marked with a dot in this drawing).

The system obviously can be extended as far as desired. The approximate gain is the sum of

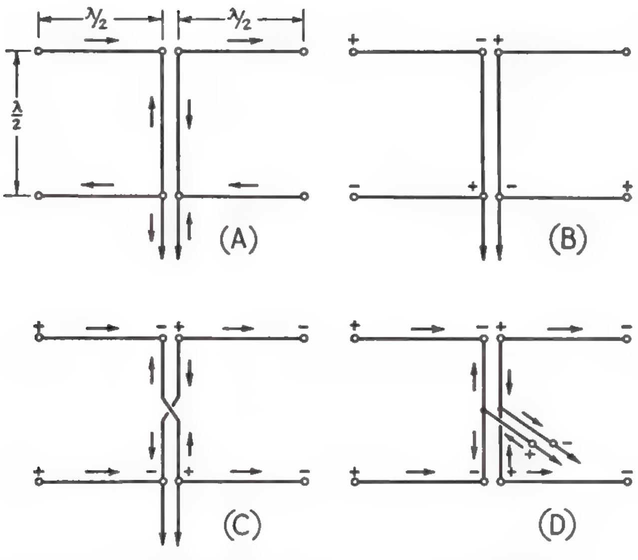


Fig. 4-41—Methods of checking the phase of currents in elements and phasing lines.

carry small currents flowing in opposite directions with respect to the center (indicated by the dot). The radiation consequently is vertically polarized. The gain is proportional to the length of the array but is somewhat smaller, because of the short radiating elements, than is obtainable from a broadside array of half-wave parallel elements of the same over-all length. The array should be 2 or more wavelengths long to secure a worthwhile gain. The system can be fed at any current loop; these occur at the centers of the vertical wires.

Another form of the Bruce array is shown at C. Because the radiating elements have twice the height, the gain is increased. The system can be fed at the center of any of the connecting lines.

the gains of one set of collinear elements and one set of broadside elements, counting the two ¼-wave sections at the ends as one element. The antenna shown, for example, is about equivalent to one set of four collinear elements and one set of two parallel broadside elements, so the total gain is approximately 4.3 + 4.0 = 8.3 db. Horizontal polarization is the only practicable type at the lower frequencies, and the lower set of elements should be at least ½ wavelength above ground for best results.

When feeding at the point shown the impedance is of the order of 600 ohms. Alternatively, this point can be closed and the system fed between any two elements, as at *X*. In this case a point near the center should be chosen so that the power distribution between elements will be as uniform as possible. The impedance at any such point will be 1000 ohms or less in systems with six or more elements.

The Bruce array is shown at B in Fig. 4-42. It consists simply of a single wire folded so that the vertical sections carry large currents in phase while the horizontal sections

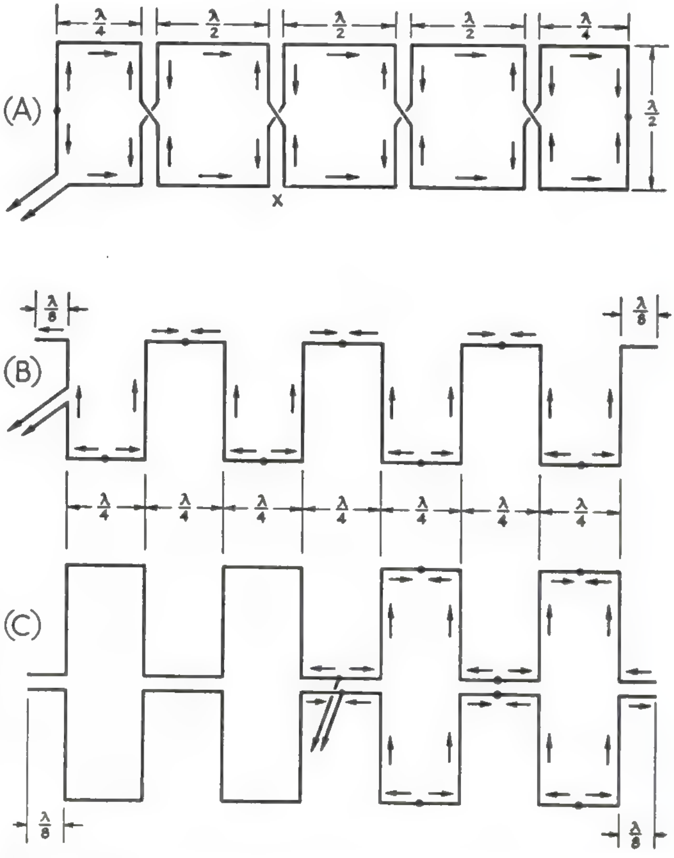


Fig. 4-42—The Sterba array (A) and two forms of the Bruce array (B and C).

Parasitic Arrays

Multielement arrays containing parasitic elements are called "parasitic" arrays even though at least one and sometimes more than one of the elements is driven. A parasitic element obtains its power through electromagnetic coupling with a driven element, as contrasted with receiving it by direct connection to the power source. A parasitic array with linear (dipole-type) elements is frequently called a "Yagi" or "Yagi-Uda" antenna after the inventors.

As explained earlier in this chapter in the section on mutual impedance, the amplitude and phase of the current induced in an antenna element depend on its tuning and the spacing between it and the driven element to which it is coupled. The fact that the relative phases of the currents in driven and parasitic elements can be adjusted is very advantageous. For example, the spacing and tuning can be adjusted to approximate the conditions that exist when two driven

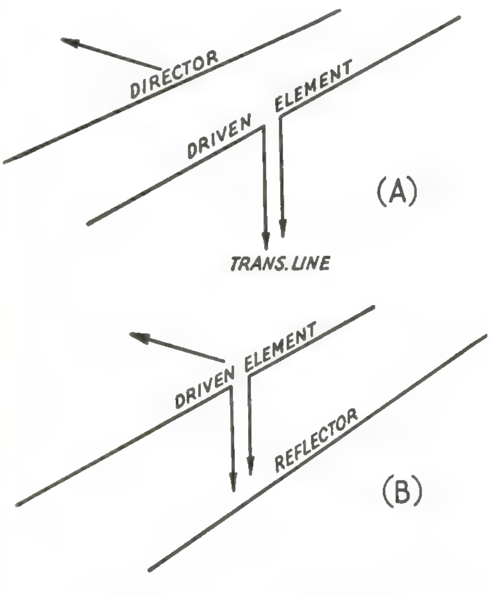


Fig. 4-43—Antenna systems using a single parasitic element. In A the parasitic element acts as a director, in B as a reflector. The arrows show the direction in which maximum radiation takes place.

elements ¼ wavelength apart are operated with a phase difference of 90 degrees (which gives a unidirectional pattern as shown in Fig. 4-27). However, complete cancellation of radiation in the rear direction is not possible when a parasitic element is used. This is because it is usually not possible to make amplitude and phase *both* reach desired values simultaneously. Nevertheless, a properly-designed parastic array can be adjusted to have a large front-to-back ratio.

The substantially unidirectional characteristic and relatively simple electrical configuration of an array using parasitic elements make it especially useful for antenna systems that are to be rotated to aim the beam in any desired direction.

Reflectors and Directors

Although there are special cases where a parasitic array will have a bidirectional (but usually not symmetrical) pattern, in most applications the pattern tends to be unidirectional. A parasitic element is called a director when it makes the radiation maximum along the perpendicular line from the driven to the parasitic element, as shown at A in Fig. 4-43. When the maximum radiation is in the opposite direction —that is, from the parasitic element through the driven element as at B—the parasitic element is called a reflector.

Whether the parasitic element operates as a director or reflector is determined by the relative phases of the currents in the driven and parasitic elements. At the element spacings commonly used (¼ wavelength or less) the current in the parasitic element will be in the right phase to make the element act as a reflector when its tuning is adjusted to the low-frequency side of resonance (inductive reactance). The parasitic element will act as a director when its tuning is adjusted to the high-frequency side of resonance (capacitive reactance). The proper tuning is ordinarily accomplished by adjusting the lengths of the parasitic elements, but the elements can be "loaded" at the center with lumped inductance or capacitance to achieve the same purpose. If the parasitic element is self-resonant the element spacing determines whether it will act as a reflector or director.

THE TWO-ELEMENT BEAM

The maximum gain theoretically obtainable with a single parasitic element, as a function of the spacing, is shown in Fig. 4-44 (from analysis by G. H. Brown). The two curves show the greatest gain to be expected when the element is tuned for optimum performance either as a director or reflector. This shift from director to reflector, with the corresponding shift in direction as shown in Fig. 4-43, is accomplished simply by tuning the parasitic element-usually, in practice, by changing its length.

With the parasitic element tuned to act as a director, maximum gain is secured when the spacing is approximately 0.1 wavelength. When the parasitic element is tuned to work as a reflector, the spacing that gives maximum gain is about 0.15 wavelength, with a fairly broad peak. The director will give slightly more gain than the reflector, but the difference is less than ½ db.

In only two cases are the gains shown in Fig. 4-44 secured when the parasitic element is self-resonant. These occur at 0.1- and 0.25-wavelength spacing, with the parasitic element acting as director and reflector, respectively. For reflector operation, it is necessary to tune the parasitic element to a lower frequency to secure maximum gain at all spacings less than

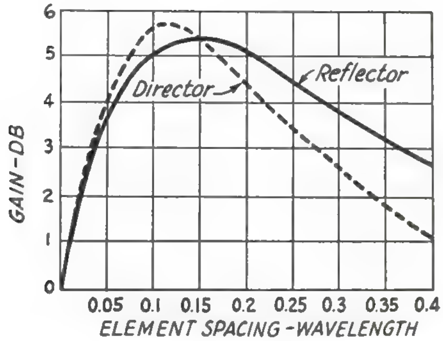


Fig. 4-44—The maximum possible gain obtainable with a parasitic element over a half-wave antenna alone, assuming that the parasitic element tuning is adjusted for greatest gain at each spacing. These curves assume no ohmic losses in the elements. In practical antennas the gain is less, particularly at close spacings.

0.25 wavelength, while at greater spacings the reverse is true. The closer the spacing the greater the detuning required. On the other hand, the director must be detuned toward a higher frequency (that is, its length must be made less than the self-resonant length) at spacings greater than 0.1 wavelength in order to secure maximum gain. The amount of detuning necessary becomes greater as the spacing is increased. At less than 0.1-wavelength spacing the director must be tuned to a lower frequency to secure the maximum gains indicated by the curve.

Input Impedance

The radiation resistance at the center of the driven element varies as shown in Fig. 4-45 for the spacings and tuning conditions that give the gains indicated by the curves of Fig. 4-44. These values, especially in the vicinity of 0.1-wavelength spacing, are quite low. The curves coincide at 0.1 wavelength, both showing a value of 14 ohms.

The low radiation resistance at the spacings giving highest gain tends to reduce the radiation efficiency. This is because, with a fixed loss resistance, more of the power supplied to the antenna is lost in heat and less is radiated, as the radiation resistance approaches the loss resistance in magnitude.

The loss resistance can be decreased by using low-resistance conductors for the antenna elements. This means, principally, large-diameter conductors, usually tubing of aluminum, copper, or copper-plated steel. Such conductors have mechanical advantages as well, in that it is relatively easy to provide adjustable sliding sections for changing length, while the fact that they can be largely self-supporting makes them well adapted for rotatable antenna construction. With half-inch or larger tubing the loss resistance in any two-element antenna should be small.

With low radiation resistance the standing waves of both current and voltage on the antenna reach considerably higher maximum values than is the case with a simple dipole. For this reason losses in insulators at the ends of the elements become more serious. The use of tubing rather than wire helps reduce the end voltage, and furthermore, the tubing does not require support at the ends, thus eliminating the insulators.

The mutual impedance between two parallel antenna elements contains reactance as well as resistance, so that the presence of a director or reflector near the driven element affects not only the radiation resistance of the driven element but also introduces a reactive component (assuming that the driven element length is such as to be resonant if the parasitic element were not there). In other words, the parasitic element detunes the driven element. The degree of detuning depends on the spacing and tuning of the parasitic element, and also on the length/diameter ratios of the elements.

With the parasitic element tuned for maximum gain, the effect of the coupled reactance is to make the driven element "look" more inductive, with the parasitic element tuned as a reflector, than it does when the parasitic element is tuned as a director. That is, the driven element should be slightly longer, if the parasitic element is a director, than when the parasitic element is a reflector. These remarks apply to spacings between about 0.1 and 0.25 wavelength, but are not necessarily true for other spacings.

Self-Resonant Parasitic Elements

The special case of the self-resonant parasitic element is of interest, since it gives a good idea

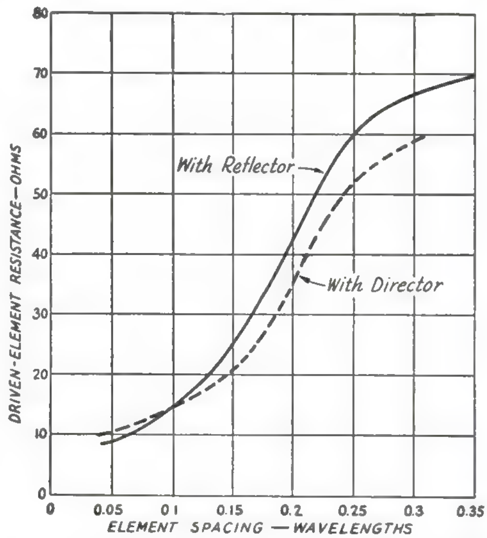


Fig. 4-45—Radiation resistance at center of driven element as a function of element spacing, when the parasitic element is adjusted for the gains given in Fig. 4-44.

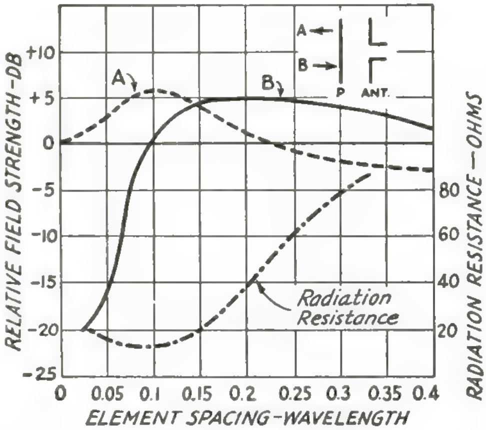


Fig. 4-46—Theoretical gain of a two-element parasitic array over a half-wave dipole as a function of element spacing when the parasitic element is self-resonant.

of the performance as a whole of two-element systems, even though the results can be modified by detuning the parasitic element. Fig. 4-46 shows gain and radiation resistance as a function of the element spacing for this case. Relative field strength in the direction A of the small drawing is indicated by Curve A; similarly for Curve B. The front-to-back ratio at any spacing is the difference between the values given by curves A and B at that spacing. Whether the parasitic element is functioning principally as a director or reflector is determined by whether Curve A or Curve B is on top; it can be seen that the principal function shifts at about 0.14-wavelength spacing. That is, at closer spacings the parasitic element is principally a director, while at greater spacings it is chiefly a reflector. At 0.14 wavelength the radiation is the same in both directions; in other words, the antenna is bidirectional with a theoretical gain of about 4 db.

The front-to-back ratios that can be secured with the parasitic element self-resonant are not very great except in the case of extremely close spacings. Spacings of the order of 0.05 wavelength are not very practicable with outdoor construction since it is difficult to make the elements sufficiently stable mechanically. Ordinary practice is to use spacings of at least 0.1 wavelength and detune the parasitic element for greatest attenuation in the backward direction.

The radiation resistance increases rapidly for spacings greater than 0.15 wavelength, while the gain, with the parasitic element acting as a reflector, decreases quite slowly. If front-to-back ratio is not an important consideration, a spacing as great as 0.25 wavelength can be used without much reduction in gain. At this spacing the radiation resistance approaches that of a half-wave antenna alone. Spacings of this order are particularly suited to antennas using wire elements, such as multielement arrays consisting of combinations of collinear and broadside elements.

Front-to-Back Ratio

The tuning conditions that give maximum gain forward do not give maximum signal reduction, or attenuation, to the rear. It is necessary to sacrifice some gain to get the highest front-to-back ratio. The reduction in backward response is brought about by adjustment of the tuning or length of the parasitic element. With a reflector, the length must be made slightly greater than that which gives maximum gain, at spacings up to 0.25 wavelength. The director must be shortened somewhat to achieve the same end, with spacings of 0.1 wavelength and more. The tuning condition, or element length, which gives maximum attenuation to the rear is considerably more critical than that for maximum gain, so that a good front-to-back ratio can be secured without sacrificing more than a small part of the gain.

For the sake of good reception, general practice is to adjust for maximum front-to-back ratio rather than for maximum gain. Larger front-to-back ratios can be secured with the

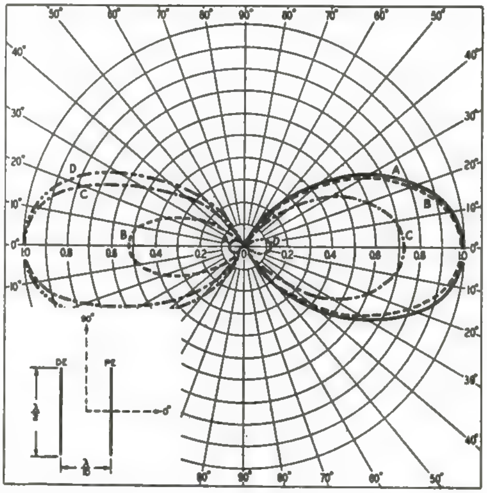


Fig. 4-47—Experimentally-determined "horizontal" directive patterns of horizontally-polarized two-element parasitic arrays at a height of 1¼ wavelengths. These patterns are for a wave angle of 12 degrees. The curves represent the following conditions, approximately:

A—Parasitic element tuned for maximum gain as a director.

B—Parasitic element self-resonant.

C—Parasitic element tuned for maximum gain as a reflector.

D—Parasitic element tuned for maximum front-to-back ratio as a reflector.

The spacing between elements is 0.1 wavelength.

The patterns should not be compared for gain, since they are plotted on a relative basis to an arbitrarily-chosen maximum of 1.0.

parasitic element operated as a director rather than as a reflector. With the optimum director spacing of 0.1 wavelength, the front-to-back ratio with the director tuning adjusted for maximum gain is only 5.5 db. (the back radiation is equal to that from the antenna alone). By proper director tuning, however, the ratio can be increased to 17 db.; the gain in the desired direction is in this case 4.5 db., or 1 db. less than the maximum obtainable.

Directional Patterns

The directional patterns obtained with two-element arrays will vary considerably with the tuning and spacing of the parasitic element. Typical patterns are shown in Figs. 4-47 and 4-48, for four cases where the parasitic element tuning or length is approximately adjusted for optimum gain as a director, for self-resonance, for optimum gain as a reflector, and for optimum front-to-back ratio as a reflector. Over this range of adjustment the width of the main beam does not change significantly. These patterns are based on experimental measurements by J. L. Gillson, W3GAU.

Bandwidth

The bandwidth of the antenna can be specified in various ways, such as the width of the band over which the gain is higher than some stated figure, the band over which at least a given front-to-back ratio is obtained, or the band over which the standing-wave ratio on the transmission line can be maintained below a chosen value. The latter is probably the most useful, since the s.w.r. not only determines the percentage power loss in the transmission line but also affects the coupling between the transmitter and the line.

The bandwidth from this latter standpoint depends on the Q of the antenna (see Chapter Three). The Q of close-spaced parasitic arrays is quite high, with the result that the frequency range over which the s.w.r. will stay below a specified maximum value is relatively narrow. The data in Table 4-I, for a driven element and close-spaced director, are from experimental measurements made by J. P. Shanklin. The antenna with 0.075-wavelength spacing will, through a suitable matching device, operate with an s.w.r. of less than 3 to 1 over a band having a width equal to about 3 per cent of the center frequency (this corresponds to the width of the 14-Mc. band for a 14-Mc. antenna, for example) and maintain a front-to-back ratio of approximately 10 db. or better over this band. At greater element spacings than those shown in the Table the Q is smaller and the bandwidth consequently greater, but the front-to-back ratio is smaller. This is to be expected from the trend shown by the curves of Fig. 4-46. The gain is practically constant at about 5 db. for all the spacings shown in Table 4-I.

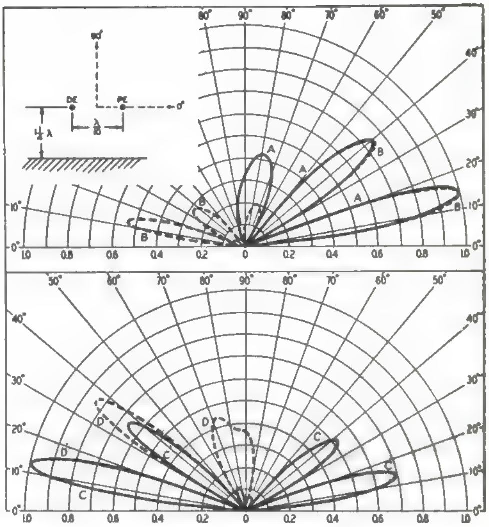


Fig. 4-48—Vertical patterns of a horizontally-polarized two-element array under the conditions given in Fig. 4-47. These patterns are in the vertical plane at right angles to the antenna elements.

The same series of experimental measurements showed that with the parasitic element tuned as a reflector for maximum front-to-back ratio the optimum spacing was 0.2 wavelength. The maximum front-to-back ratio was determined to be 16 db. In both the director and reflector cases the front-to-back ratio decreased rather rapidly as the operating frequency was moved away from the frequency for which the system was tuned. With the reflector at 0.2 wavelength spacing and tuned for maximum front-to-back ratio the input resistance was found to be 72 ohms and the Q of the antenna was 4.7.

The antenna elements used in these measurements had a length/diameter ratio of 330. A smaller length/diameter ratio will decrease the rate of reactance change with length and hence decrease the Q, while a larger ratio will increase the Q. The use of fairly thick elements is desirable when maximum bandwidth is sought.

THE THREE-ELEMENT BEAM

It is readily possible to use more than one parasitic element in conjunction with a single

TABLE 4-I

Feed Impedance and Front-to-Back Ratio of a Fed Dipole with One Director

Element Spacing	*Fed Dipole Length*	*Director Length*	*Input Resistance at Band Center*	*Q*	*Front-to-Back Ratio at Band Center*
0.050 λ	0.509 λ	0.484 λ	13.2 ohms	53.2	20 (26 db.)
0.075	0.504	0.476	24.4	29.4	8.3 (18 db.)
0.100	0.504	0.469	28.1	20.0	4.3 (12.7 db.)

driven element. With two parasitic elements the optimum gain and directivity result when one is used as a reflector and the second as a director. Such an antenna is shown in Fig. 4-49.

As the number of parasitic elements is increased, the problem of determining the optimum element spacings and lengths to meet given specifications—i.e., maximum gain, maximum front-to-back ratio, maximum bandwidth, and so on—becomes extremely tedious because of the large number of variables. In general, it can be said that when one of these quantities—gain, front-to-back ratio, or bandwidth—is maximized the other two cannot be. Also, if it is desired to design the antenna to have a specific

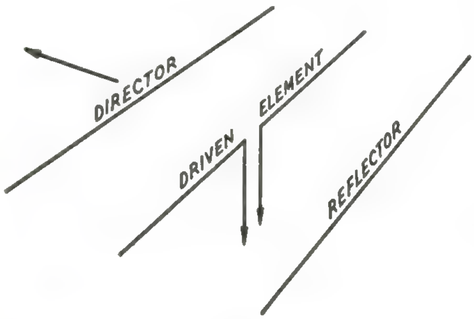


Fig. 4-49—Antenna system using a driven element and two parasitic elements, one as a reflector and one as a director.

input impedance for matching a transmission line, the other three cannot be maximized.

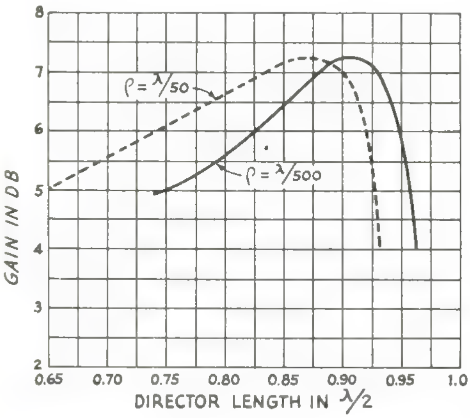


Fig. 4-50—Gain of a 3-element Yagi over a dipole as a function of the director length, for 0.2 wavelength spacing between driven element and director and the same spacing between driven element and reflector. These curves show how the element thickness affects the optimum length. ρ is the element radius expressed as a fraction of the wavelength. $\lambda/500$ corresponds to an element diameter of approximately 1 inch at 50 Mc. Where the fractional-wavelength radius is smaller, as on the lower frequencies, the optimum director length will be somewhat greater.

Power Gain

A theoretical investigation of the 3-element case (director, driven element and reflector) has indicated a maximum gain of slightly more than 7 db. (Uda and Mushiake). A number of experimental investigations have shown that the optimum spacing between the driven element and reflector is in the region of 0.15 to 0.25 wavelength, with 0.2 wavelength representing probably the best over-all choice. With 0.2 wavelength reflector spacing, Fig. 4-50 shows the variation in gain with director length, with the director also spaced 0.2 wavelength from the driven element, and Fig. 4-51 shows the gain variation with director spacing. (These curves are from work by Carl Greenblum.) It is obvious that the director spacing is not especially critical, and that the over-all length of the array (boom length in the case of a rotatable antenna) can be anywhere between 0.35 and 0.45 wavelength with no appreciable difference in gain.

Wide spacing of both elements is desirable not only because it results in high gain but also because adjustment of tuning or element length is less critical and the input resistance of the driven element is higher than with close spacing. The latter feature improves the efficiency of the antenna and makes a greater bandwidth possible. However, a total antenna length, director to reflector, of more than 0.3 wavelength at frequencies of the order of 14 Mc. introduces considerable difficulty from a constructional standpoint, so lengths of 0.25 to 0.3 wavelength are frequently used for this band, even though they are less than optimum.

In general, the gain of the antenna drops off less rapidly when the reflector length is increased beyond the optimum value than it does for a corresponding decrease below the optimum value. The opposite is true of a director, as

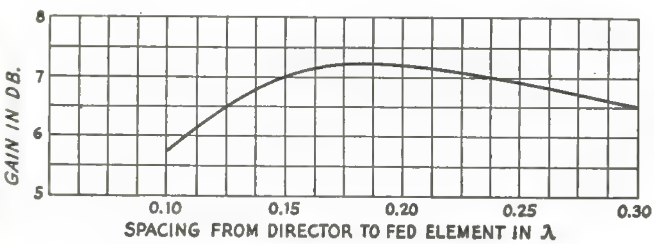


Fig. 4-51—Gain of 3-element Yagi versus director spacing, the reflector spacing being fixed at 0.2 wavelength.

shown by Fig. 4-50. It is therefore advisable to err, if necessary, on the long side for a reflector and on the short side for a director. This also tends to make the antenna performance less dependent on the exact frequency at which it is operated, because an increase above the design frequency has the same effect as increasing the length of both parasitic elements, while a decrease in frequency has the same effect as shortening both elements. By making the director slightly short and the reflector slightly long, there will be a greater spread between the upper and lower frequencies at which the gain starts to show a rapid decrease.

Input Impedance

The radiation resistance as measured at the center of the driven element of a 3-element array can vary over a fairly wide range since it is a function of the spacing and tuning of the parasitic elements. There are, however, certain fairly well-defined trends. (1) The resistance tends to reach a minimum at the parasitic-element tuning condition that gives maximum gain, becoming larger as the element is detuned in either direction—that is, made longer or shorter. (2) The resistance tends to be lower the closer the spacing between the parasitic and driven elements. Values of the order of 10 ohms are typical with a 3-element beam having 0.1 wavelength director spacing, when the director length is adjusted for maximum gain. This can be raised considerably—to 50 ohms or more—by sufficient change in director length, at a sacrifice of gain. The minimum value of resistance increases with increased director spacing, and is of the order of 30 ohms at a spacing of 0.25 wavelength.

As in the case of the two-element beam, tuning and spacing of the parasitic elements affect the reactance of the driven element; that is, a change in the spacing or length of the parasitic elements will tend to change the resonant frequency of the driven element. It is generally found, however, that the resonant length of the driven element with the parasitic elements properly tuned does not differ greatly from its resonant length with the parasitic elements removed. This is because the two parasitic elements reflect opposite kinds of reactance into the driven element and hence tend to cancel each other's effects in this respect.

Fig. 4-52 shows the results of experimental measurements made by J. P. Shanklin of the input resistance of 3-element arrays having an over-all length (director to reflector) of 0.3 wavelength. The curves gives resistance contours as a function of the spacing between the driven element and the reflector and the length of the director. So long as the reflector length was in the optimum region for good front-to-back ratio, as described in the next section, small changes in reflector length were found to have only a comparatively small effect on the input resistance. In using Fig. 4-52, it is to be understood that the spacing between the director and driven element is equal to the difference between 0.3 wavelength and the selected driven-element-to-reflector spacing, since the length of the array was held constant.

The elements used in obtaining the data in Fig. 4-52 had a length/diameter ratio of 330.

Front-to-Back Ratio

The element lengths and spacings are more critical when a high front-to-back ratio is the objective than when the antenna is designed for maximum gain. Some gain must be sacrificed for the sake of a good front-to-back ratio, just as in the case of the two-element array. In general, a high front-to-back ratio requires fairly close spacing between the director and driven element, but considerably larger spacings are optimum for the reflector.

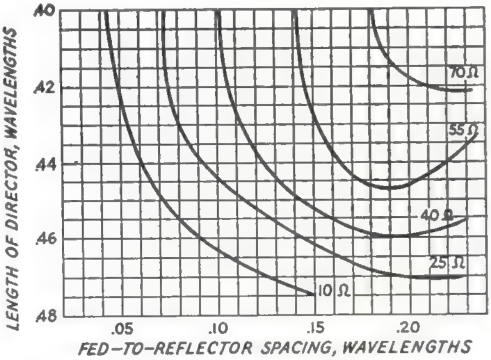


Fig. 4-52—Resonant resistance of fed dipole in a 3-element parasitic antenna, overall length 0.3 wavelength.

The front-to-back ratio will change more rapidly than the gain when the operating frequency differs from that for which the antenna was adjusted.

The front-to-back ratio tends to decrease with increased spacing between the elements. However, with a director spacing of about 0.2 wavelength it is possible to secure a very good front-to-*side* ratio, which may be a useful feature in some locations. As in the case of front-to-back ratio, the reflector spacing has a considerably lesser effect.

Fig. 4-53 is a chart showing the spacing and director tuning region for good front-to-back ratio. It is based on the same experimental data as Fig. 4-52, using an antenna having an over-all length of 0.3 wavelength and an element length/diameter ratio of 330. The shaded area marked "14 Mc. to 14.4 Mc." gives the range of spacings and corresponding limits of director lengths for which the front-to-back ratio will be higher than 12 db. over the entire band of frequencies. The shaded area marked "27 Mc. to 29.7 Mc." gives similar data for the same performance over the latter band. Combina-

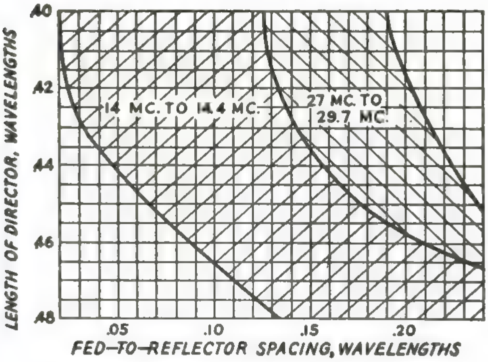


Fig. 4-53—Effect of director length and spacing between fed element and reflector on front-to-back ratio, three-element antenna. Shaded areas are for more than 4-to-1 voltage front-to-back ratio over the frequency ranges designated.

In the case of the 14-Mc. range the front-to-back ratio is greater than 4 to 1 anywhere to the right of the curve bounding the left-hand side of the shaded area. The bandwidth as defined by this limiting front-to-back ratio becomes greater in the upper right-hand portion of the figure (wide-band region), and less in the lower portion of the shaded area (narrow-band region).

The 27-29.7 Mc. range has the greatest bandwidth approximately midway between the two boundary curves.

tions of spacing and director length that fall near the top of the chart give greater bandwidth, in this respect, than the combinations that fall near the bottom. (For any selected combination, the corresponding input resistance of the driven element may be found from Fig. 4-52.)

The antenna gains found in these measurements varied from about 4 to 6 db., the larger gains being associated with the narrower-band spacing-length combinations represented by the lower portions of the shaded areas. These gains compare with the approximately 7 db. obtainable with wider spacings when the antenna is adjusted for maximum gain, with front-to-back ratio a secondary consideration.

Bandwidth

The bandwidth with respect to input impedance, as evidenced by the change in standing-wave ratio over a band of frequencies, will in general be smaller the smaller the input resistance. This in turn becomes smaller when the spacing between elements is decreased. Hence close spacings are usually associated with small band widths, especially when the element lengths are adjusted for maximum gain.

Fig. 4-54, also from data obtained in the experimental measurements of J. P. Shanklin, shows how the Q of a 3-element antenna having a total length of 0.3 wavelength varies as a function of spacing and director length. The data are for an element length/diameter ratio of 330, but should hold sufficiently well for ratios between 200 and 400. From the standpoint of impedance bandwidth, the upper right-hand region of the chart is best since this region is associated with low values of Q.

In these measurements it was found that the length of the reflector for optimum front-to-back ratio did not vary over much of a range. In the "narrow-band" region of Figs. 4-53 and 4-54 it was 0.51 wavelength, increasing to 0.525 wavelength in the "wide-band" region. The proper driven-element length was found to be 0.49 wavelength (at the center of the band) for all conditions.

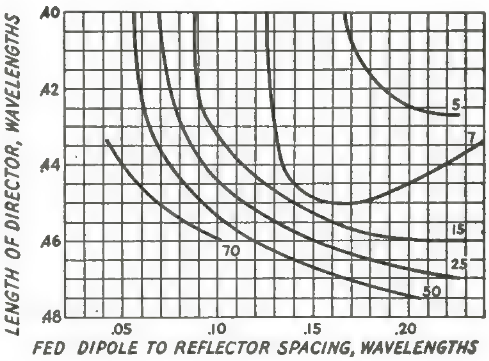


Fig. 4-54—Q of input impedance of fed dipole in a 3-element parasitic antenna, over-all length 0.3 wave length.

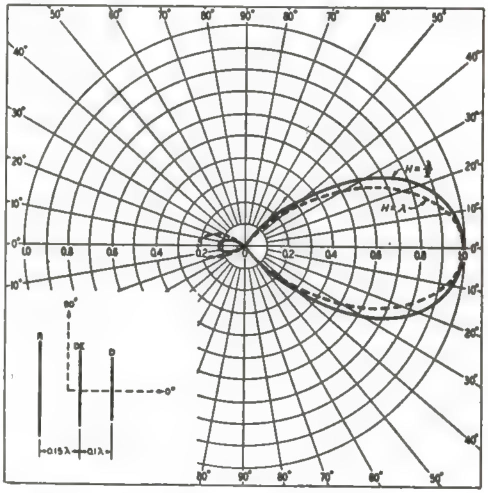


Fig. 4-55—Measured radiation patterns of a horizontally-polarized 3-element array having the director 0.1 wavelength and the reflector 0.15 wavelength from the driven element. Element tuning adjusted for maximum gain. These "horizontal" patterns are for the wave angles at which the lowest lobe (Fig. 4-56) has its maximum. The wave angles are 28 degrees at a height of ½ wavelength and 12 degrees at a height of 1 wavelength.

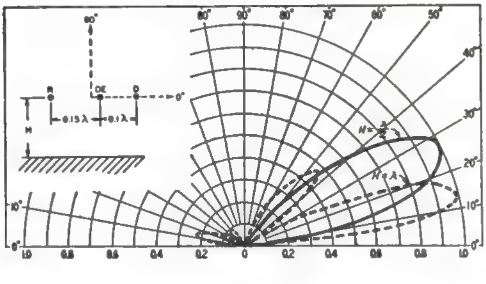


Fig. 4-56—Vertical patterns of the antenna of Fig. 4-55 in the vertical plane at right angles to the direction of the antenna elements.

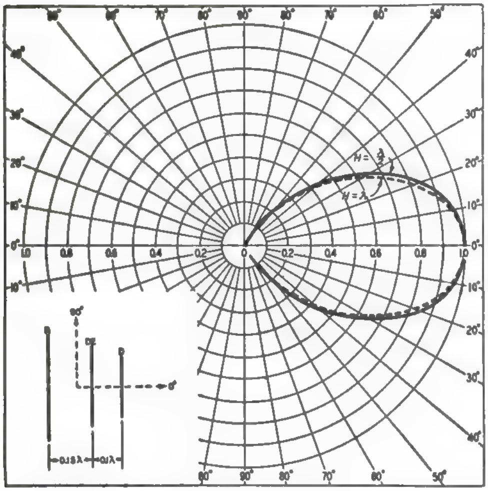


Fig. 4-57—Measured radiation patterns of a horizontally-polarized 3-element parasitic array having the director 0.1 wavelength from the driven element and the reflector 0.15 wavelength from the driven element. Element tuning adjusted for maximum front-to-back ratio. Heights and wave angles are same as in Fig. 4-55.

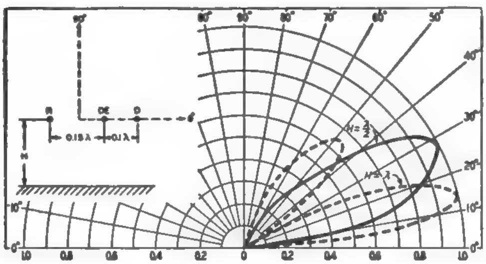


Fig. 4-58—Vertical patterns of the antenna of Fig. 4-57 in the vertical plane at right angles to the direction of the antenna elements.

Similar tests were made on antennas having over-all lengths of 0.2 and 0.4 wavelengths. The conclusion was (1) that the smaller length would give high front-to-back ratios but with high Q values and consequently small bandwidth; (2) with 0.4 wavelength spacing the Q values were low enough for good bandwidth but the front-to-back ratio was smaller.

The Q values given by the chart can be used as described in Chapter Three to find the bandwidth over which the s.w.r. will not exceed a specified value.

The low values of radiation resistance are accompanied by a high degree of selectivity in the antenna; that is, its impedance is constant over only a small frequency range. These changes in impedance make it troublesome to couple power from the transmitter to the line. Such difficulties can be reduced by using wider spacing—in particular, using spacings of the order of 0.2 wavelength or more.

Directive Patterns

The directive patterns of Figs. 4-55 to 4-58 inclusive are, like those of Figs. 4-47 and 4-48, based on experimental measurements made by W3GAU at v.h.f. They show that the beam is somewhat sharper, as is to be expected, when the parasitic-element tuning is adjusted for maximum gain. Increasing the height of the antenna will of course lower the wave angle since the shape and amplitude of the vertical lobes are determined by the ground-reflection factors given in Chapter Two as well as by the free-space pattern of the antenna itself.

FOUR-ELEMENT ARRAYS

Parasitic arrays having a driven element and three parasitic elements—reflector and two directors—are frequently used at the higher frequencies, 28 Mc. and up. This type of antenna is shown in Fig. 4-59.

Close spacing is undesirable in a four-element antenna because of the low radiation resistance. An optimum design, based on an experimental determination at 50 Mc., uses the following spacings:

Driven element to reflector—0.2 wavelength
Driven element to first director—0.2 wavelength
First director to second director—0.25 wavelength

Using a length/diameter ratio of about 100 for the elements, the element lengths for maximum gain were found to be

Reflector—0.51 wavelength
Driven Element—0.47 wavelength
First Director—0.45 wavelength
Second Director—0.44 wavelength

The input resistance with the above spacings and dimensions was of the order of 30 ohms and the antenna gave useful gain over a total bandwidth equal to about 4 per cent of the center frequency.

LONG YAGIS

Parasitic arrays are not limited as to the number of elements that can be used, although it is hardly practical to use more than four at frequencies below 30 Mc. However, on the v.h.f.

TABLE 4-II

Optimum Element Spacings for Multielement Yagi Arrays DE — driven element; R — reflector; D — director.

No. Elements	R-DE	DE-D_1	D_1-D_2	D_2-D_3	D_3-D_4	D_4-D_5	D_5-D_6
2	0.15λ–0.2λ						
2		0.07λ–0.11λ					
3	0.16 –0.23	0.16 –0.19					
4	0.18 –0.22	0.13 –0.17	0.14λ–0.18λ				
5	0.18 –0.22	0.14 –0.17		0.17λ–0.23λ			
6	0.16 –0.20	0.14 –0.17	0.16 –0.25	0.22 –0.30	0.25λ–0.32λ		
8	0.16 –0.20	0.14 –0.16	0.18 –0.25	0.25 –0.35	0.27 –0.32	0.27λ–0.33λ	0.30λ–0.40λ
8 to N	0.16 –0.20	0.14 –0.16	0.18 –0.25	0.25 –0.35	0.27 –0.32	0.27 –0.33	0.35 –0.42

N = any number; director spacings beyond D_6 should be 0.35–0.42λ.

bands an array that is long in terms of wavelength is often of practicable physical size. Several independent investigations of the properties of multielement Yagi antennas have shown that in a general way the gain of the

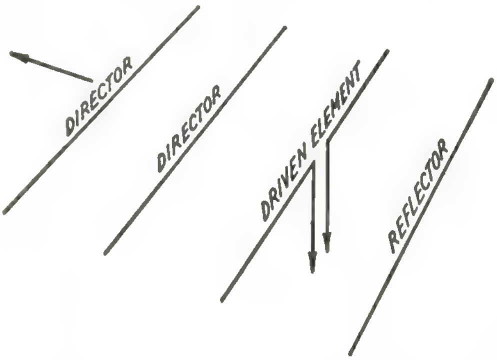


Fig. 4-59—A "four-element" antenna system, using two directors and one reflector in conjunction with a driven element.

antenna expressed as a power ratio is proportional to the length of the array, provided the number, lengths and spacings of the elements are properly chosen.

The results of one such study (by Carl Greenblum) are shown in terms of the number of elements in the antenna in Figs. 4-60 and 4-61. In every case the antenna consists of a driven element, one reflector, and a series of directors properly spaced and tuned. Thus if the antenna is to have a gain of 12 db., Fig. 4-60 shows that 8 elements—driven, reflector, and six directors—will be required, and Fig. 4-61 shows that for such an 8-element antenna the array length required is 1.75 wavelength.

Table 4-II shows the optimum element spacings determined from the Greenblum investigations. There is a fair amount of latitude in the placement of the elements along the length of the array, although the optimum tuning of the element will vary somewhat with the exact spacing chosen. Within the spacing ranges shown, the gain will not vary more than 1 db. provided the director lengths are suitably adjusted.

The optimum director lengths are in general greater the closer the particular director is to the driven element, but the length does not uniformly decrease with increasing distance from the driven element. Fig. 4-62 shows the experimentally-determined lengths for various element diameters, based on cylindrical elements supported by mounting through a cylindrical metal boom two or three times the element diameter. The curves probably would not be useful for other shapes.

In another study of long Yagi antennas at v.h.f., J. A. Kmosko, W2NLY, and H. G. Johnson, W6QKI, reached essentially the same general conclusions concerning the relationship between overall antenna length and power gain, although their gain figures differ from those of Greenblum. The comparison is shown in Fig. 4-63. The Kmosko-Johnson results are based on a somewhat different element spacing and a construction in which thin director elements are supported above the metal boom rather than running through it. In their optimum design the first director is

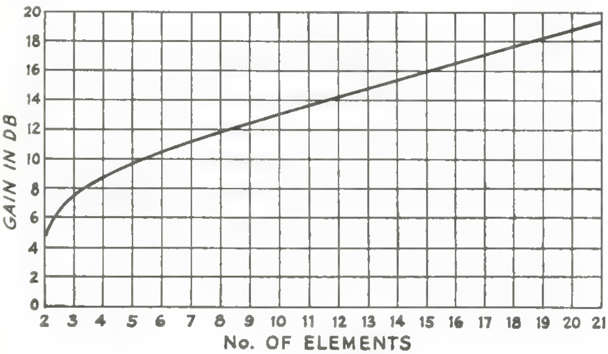


Fig. 4-60—Gain in db. over a half-wave dipole vs. the number of elements of the Yagi array, assuming the array length is as given in Fig. 4-61.

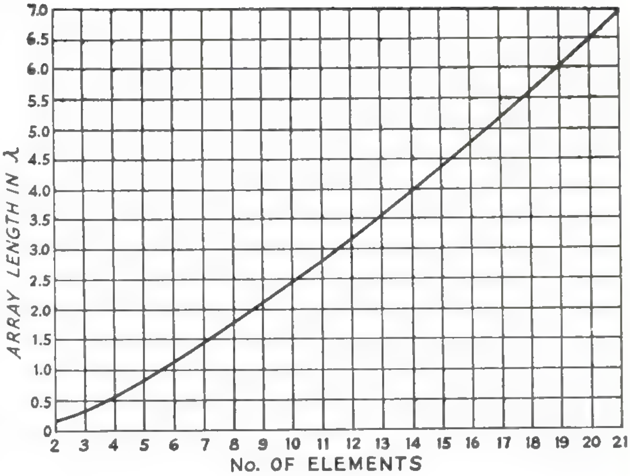


Fig. 4-61—Optimum length of Yagi antenna as a function of number of elements.

spaced 0.1 wavelength from the driven element. The next two directors are slightly over 0.1 wavelength apart, the fourth director is approximately 0.2 wavelength from the third, and succeeding directors are spaced 0.4 wavelength apart. A practical design for such an antenna for 144-Mc. is shown in Chapter Ten. The Kmosko-Johnson figures are based on a simplified method of computing gain from the beam width of the antenna pattern, the beam widths having been measured experimentally. The Greenblum data is from experimental measurement of gain.

Experimental gain figures based on measurements made at 3.3-centimeter wavelength by H. W. Ehrenspeck and H. Poehler, shown by a third curve in Fig. 4-63, indicate lower gain for a given antenna length but confirm the gain-*vs.*-length trend. These measurements were made over a large ground plane using elements of the order of one-quarter wavelength high. The general conclusions of this study were 1) that the reflector spacing and tuning is independent of the other antenna dimensions, the optimum fed-element to reflector spacing being in the neighborhood of 0.25 wavelength but not critical; (2) for a given antenna length the gain is practically independent of the number of directors provided the director-to-director spacing does not exceed 0.4 wavelength; 3) that the optimum director tuning differs with different director spacings but that for constant spacings all directors can be similarly tuned; and 4) a slight improvement in gain results from using an extra director spaced about 0.1 wavelength from the driven element.

The agreement between these three sets of measurements is not as close as might be wished for, which simply confirms the difficulty of determining optimum design where a multiplicity of elements is used, and of measuring gain with a degree of accuracy that will permit reconciliation of the results obtained by various observers. There is, however, agreement on the general principle that length is of greater importance than the number of elements, within the limit of a maximum element spacing of 0.4 wavelength.

It is an interesting fact that the feed-point impedance and bandwidth of long Yagis depends almost entirely on the two or three parasitic elements closest to the driven element, presumably because those farther from the driven element are relatively loosely coupled to it. In this respect, therefore, the information already given in connection with three-element arrays is quite applicable.

STACKED YAGIS

Parasitic arrays can be stacked either in broad-

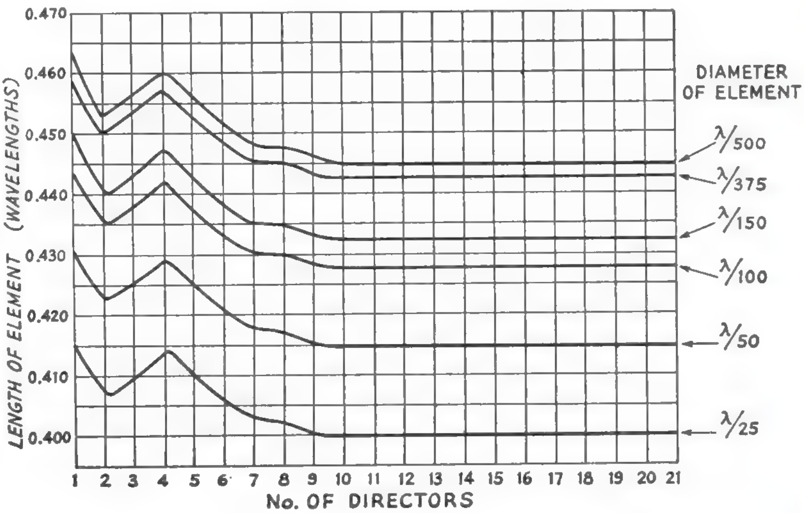


Fig. 4-62—Length of director vs its position in the array, for various element thicknesses.

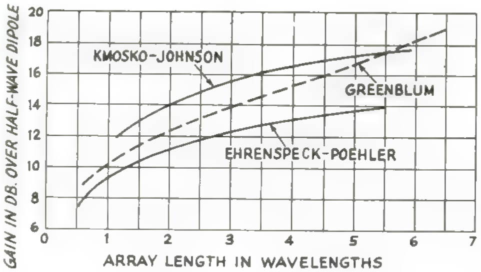


Fig. 4-63—Gain of long Yagi antennas as a function of over-all length. The antenna consists of a driven element, a single reflector spaced approximately one-fourth wavelength from the driven element, and a series of directors spaced as described in the text. The three curves represent the results of three independent studies.

side or collinear fashion for additional directivty and gain. The increase in gain that can be realized is dependent on the spacing between the individual arrays. It is assumed, of course, that all the individual arrays making up the stacked system are identical, and that in the case of broadside stacking the corresponding elements are parallel and lie in planes perpendicular to the axis of the individual arrays. In collinear stacking, it is assumed that the corresponding elements are collinear and all elements of the individual arrays lie in the same plane. In both cases the driven elements must be fed in phase.

The decrease in beamwidth of the main radiation lobe that accompanies stacking is, in the general case, accompanied by a splitting off of one or more sets of side lobes. These will have an amplitude depending on the shape of the directive pattern of the unit array, the number of unit arrays, and their spacing. An optimum spacing is one which gives as much gain as possible on the condition that these side lobes do not exceed some specified amplitude relative to the main lobe. Fig. 4-64 shows the optimum spacings for three such conditions (no side lobes, side lobes down 10 db., and side lobes down 20 db.) as a function of the half-power beamwidth of the unit array, from calculations by H. W. Kaspar, K2GAL. Maximum gain occurs when the side lobes are approximately 10 db. down, as indicated in the figure.

A single 3-element array will have a half-power beamwidth (free space) of approximately 75 degrees, and from Fig. 4-64 it can be determined that the optimum spacing for maximum gain will be slightly over ¾ wavelength. Measurements by Greenblum have shown that the stacking gain that can be realized with two such Yagi antennas is approximately 3 db., remaining practically constant at spacings from ¾ to 2 wavelengths, but decreasing rapidly at spacings less than ¾ wavelength. With spacings less than about ½ wavelength stacking does not give enough gain to make the construction of a stacked array worthwhile.

If reduction of side lobe amplitude is the principal consideration rather than gain, smaller spacings are optimum, as shown by the curves.

A similar set of curves for four stacked unit arrays is given in Fig. 4-65.

The spacings in Figs. 4-64 and 4-65 are measured between the array centers. When the unit arrays are stacked in a collinear arrangement, a spacing of less than ½ wavelength is physically impossible with full-size elements, since at ½-wavelength spacing the ends of the collinear elements will be practically touching.

FEEDING AND ADJUSTMENT

The problems of matching and adjusting parasitic arrays for maximum performance are the same in principle as with other antenna systems. Adjustment of element lengths for optimum performance usually necessitates measurements of relative field strength. However, the experience of a great many amateurs who have followed the rather laborious procedure of adjusting each element a little at a time, measuring the relative field after each such change, has accumulated a large amount of data on optimum lengths. Depending on the objective in designing the antenna—i.e., maximum gain, maximum front-to-back ratio, etc. —it is possible to predetermine the actual element lengths for a given center frequency and thus avoid the

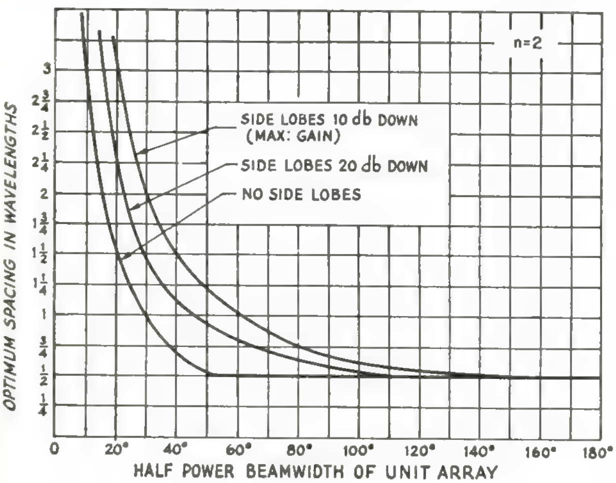


Fig. 4-64—Optimum stacking spacing for two unit arrays. The spacing for no side lobes, especially for small beam widths, may result in almost no gain improvement with stacking.

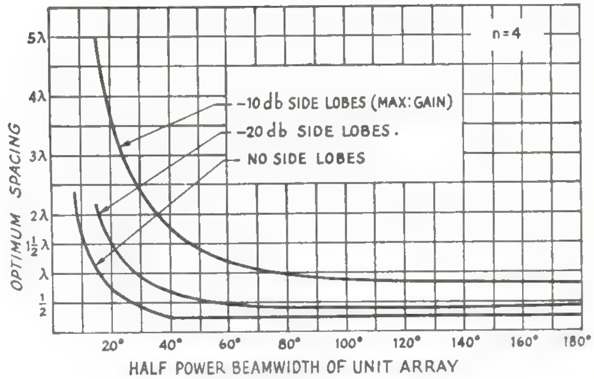


Fig. 4-65—Optimum stacking spacing for four unit arrays.

necessity for such adjustments. Charts giving proper element lengths for 3-element beams are shown in Chapter Twelve, for the maximum-gain condition, and data for maximum front-to-back ratio were given earlier in this chapter. The principal adjustment that actually needs to be made is to match the antenna to the transmission line so the standing-wave ratio on the latter is minimized.

Methods of Feed

The driven element in a parasitic array is a load for the transmission line in the same way that a driven element in any antenna system is such a load. It differs from the load presented by a simple dipole only in that the resistance may be quite low, especially if close spacings are used between elements, and the rate of change of reactance as the operating frequency is moved away from the design frequency may be greater. With low input resistance, a fairly large impedance step-up is required for matching practicable lines, and the amount of mismatch will increase more rapidly than with a simple dipole when the applied frequency is varied from that at which the line is matched.

Practically any of the matching systems detailed in Chapter Three are applicable. For example, an open quarter-wave matching section (Fig. 3-68A) can be used if the transmission line is to be 300- to 600-ohm parallel-conductor line. The quarter-wave transformer (Q) method also can be used, as shown in Fig. 3-55, with 75-ohm Twin-Lead for the transformer. This particular value of Z_0 will match a driven element having a resistive impedance of 9 to 10 ohms to a 600-ohm open-wire line, and will result in only a 2-to-1 mismatch if the driven-element impedance is as low as 4 or 5 ohms or as high as 18 to 20 ohms. The loss in a quarter-wave section of 75-ohm transmitting-type Twin-Lead, even though the s.w.r. in the matching transformer is rather high, will not be of any real consequence. Alternatively, the matching transformer may be made of 50- or 75-ohm coaxial cable. However, in such case some method of line balancing (such as those shown in Fig. 3-54) should be used. The delta match (Fig. 3-57) also may be used.

The folded dipole used as the driven element also furnishes a useful method of transforming a low antenna resistance to a value suitable for matching a transmission line. Design details are given in Chapter Three.

The choice of a matching system is affected by constructional considerations, since parasitic arrays are usually built to be rotated in operation. The T-match (Fig. 3-62) and Gamma match (Fig. 3-63) are favorite with many amateurs because they fit in well, constructionally, when the driven element is made of tubing. Another matching system that is useful when the antenna is to be continuously rotatable is the inductive coupling shown in Fig. 3-74.

The adjustment procedure for matching systems is described in detail in Chapter Three.

Broadening the Response

It has already been pointed out that the tuning conditions giving maximum gain with parasitic elements are not highly critical. However the varying amounts of reactance coupled into the driven element, as well as the fact that the radiation resistance at the center of the driven element is often very low, cause the impedance to change rapidly when the applied frequency is varied above or below the design frequency.

This impedance change can be made less rapid by using fairly wide spacing between elements, as already mentioned. It is also beneficial to use elements having a fairly large ratio of diameter to length because, as explained in Chapter Two, the impedance change with frequency is reduced when the antenna conductor has a large diameter.

The use of a folded-dipole driven element is beneficial in broadening the frequency characteristic of the antenna because of the smaller effective length/diameter ratio that results from using two or more conductors instead of one.

Adjusting Parasitic Arrays

There are two separate processes in adjusting an array with parasitic elements. One is the determination of the optimum element lengths, depending on whether maximum gain or maximum front-to-back ratio is desired. The other is matching the antenna to the transmission line. The second is usually dependent on the first, and the results observed on adjusting the element tuning may well be meaningless unless the line is equally well matched under all tuning conditions.

As stated earlier, the element tuning for maximum gain is not excessively critical, and the dimensions given by the following formulas

have been found to work well in practice for 3-element antennas:

$$\textbf{Driven element: } \textit{Length} \text{ (ft.)} = \frac{475}{f \text{ (Mc.)}}$$

$$\textbf{Director: } \textit{Length} \text{ (ft.)} = \frac{455}{f \text{ (Mc.)}}$$

$$\textbf{Reflector: } \textit{Length} \text{ (ft.)} = \frac{500}{f \text{ (Mc.)}}$$

These are average lengths determined experimentally for elements having a length/diameter ratio of 200 to 400, and with element spacings from 0.1 to 0.2 wavelength.

Many amateurs have found that very satisfactory results are secured simply by cutting the elements to the lengths given by these formulas. It has been a rather common experience that, after a considerable amount of time has been spent in trying all possible adjustments, the dimensions finally determined to be optimum are very close to those given by the formulas above or the charts in Chapter Twelve (the difference between the charts and the formulas above amount to only about one per cent in most cases), and the actual difference in gain is negligible. It appears safe to say, therefore, that in the average case there is probably little to be realized, in the way of increased gain, by spending much time in adjusting element lengths. The front-to-back ratio can often be improved, however, since it depends much more on the exact element tuning. In general, the reflector tuning is the more critical.

If the array is put up by formula, the only adjustment that need be made is to match the driven element to the transmission line. The adjustment procedure for each type of matching arrangement is described in Chapter Three.

Test Set-up

The only practicable method of adjusting parasitic element lengths for best performance is to measure the field strength from the antenna as adjustments are made. Measurements on a relative basis are entirely satisfactory for the purpose of determining the operating conditions that result in the maximum output or greatest front-to-back ratio. For this purpose the measuring equipment does not need to be calibrated; the only requirement is that it indicate whether the signal is stronger or weaker.

If the help of a nearby amateur owning a receiver with an S-meter can be enlisted, the S-meter indications can be used to indicate the relative field strength. A few precautions must be taken if this method is to be reliable. The receiving antenna must have the same polarization as the transmitting antenna under test (this is usually horizontal), and should be reasonably high above its surroundings. The receiving system should be checked for pick-up on the transmission line to make sure that the indications given by the receiver are caused entirely by energy picked up by the receiving antenna itself. This can be checked by temporarily disconnecting the line from the antenna (but leaving it in place) and observing the signal strength on the S-meter. If the reading is not several S points below the reading with the antenna connected the readings cannot be relied upon when adjusting the transmitting antenna for maximum gain. In checking the front-to-back ratio, the stray pickup at the receiving installation must be well below the smallest signal received via the antenna, if the adjustments are to mean anything at all.

Another method of checking field strength is to use a field-strength indicator of the crystal-detector type. The preferable method of using such an indicator is to connect it to a dipole antenna mounted some distance away and at a height at least equal to that of the transmitting antenna. There should be no obstructions between the two antennas, and both should have the same polarization. The receiving dipole need not be a half-wave long, although that length is desirable because it will increase the ratio of energy picked up on the antenna to energy picked up by stray means. To prevent coupling effects the distance between the two antennas should be at least three wavelengths. At shorter distances the mutual impedance may be large enough to cause the receiving antenna to tend to become part of the transmitting system, which can lead to false results. A recommended type of indicating system is shown in Fig. 4-66. The transmission line should drop vertically down to the indicator, to avoid stray pick up. This pick-up can be checked as described in the preceding paragraph. If the distance between the two antennas is such that greater sensitivity is needed a reflector may be placed ¼ wavelength behind the receiving dipole.

Adjustment Procedure

It is advisable first to set the element lengths to those given by the formulas and then match the driven element to the transmission line obtaining as low an s.w.r. as possible. In subsequent adjustments a close watch should be kept on the s.w.r. and the transmitter power input should be maintained at exactly the same figure throughout. If the s.w.r. changes enough to affect the coupling at the transmitter when an adjustment is made, but not enough to raise the line loss significantly (see Fig. 3-18), readjust the coupling to bring the input back to the same value. If the line loss increases more than a fraction of a decibel, *rematch at the driven element.* If this is not done, the results may be entirely misleading; it is absolutely necessary to maintain constant power input to the driven element if adjustment of directors or reflectors is to give meaningful results.

The experience of most amateurs in adjusting parasitic arrays indicates that there is not a great deal of preference in the order in which elements are tuned, but that there is slightly less

interlocking if the director is first adjusted to give maximum gain and the reflector is then adjusted to give either maximum gain or maximum front-to-back ratio, whichever is desired. After the second parasitic element has been adjusted, go back and check the tuning of the first to make sure that it has not been thrown out of adjustment by the mutual coupling. If there are three parasitic elements, the other two should be checked each time an appreciable change is made in one. The actual lengths should not be very far from those given by the formulas when the optimum settings are finally determined. As already pointed out, the reflector length may be somewhat greater when adjusted to give maximum front-to-back ratio.

Radiation from the transmission line must be eliminated, or at least reduced to a very low value compared with the radiation from the antenna itself, if errors are to be avoided. Conditions are usually favorable to low line radiation in horizontally-polarized rotatable parasitic arrays because the line is usually symmetrical with respect to the antenna and is brought away perpendicular to it, at least for a half wave length or so. Nevertheless the line radiation can be appreciable unless the line is detuned as described in Chapter Three. With coaxial line some method of line balancing at the antenna always should be incorporated.

After arriving at the optimum adjustments at the frequency for which the antenna was designed, the performance should be checked over a frequency range either side of the design frequency to observe the sharpness of response. If the field strength falls off rapidly with frequency, it may be desirable to shorten the director a bit to increase the gain at frequencies above resonance and lengthen the reflector slightly to increase it at frequencies below resonance. Do not confuse the change in s.w.r. with the change in antenna gain. The antenna itself may give good gain over a considerable frequency range, but the s.w.r. may vary between wide limits in this range. To check the *antenna* behavior, keep the power input to the transmission line constant and rematch the driven element to the line, as suggested above, whenever the line losses increase appreciably. If such rematching is found necessary over the band of frequencies to be used, it may be advisable to retune the system to give a higher input resistance and thus decrease the selectivity, even though some gain is sacrificed in so doing.

Adjustment by Reception

As an alternative to applying power to the array and checking the field strength, it is possible to adjust the array by measuring received signal strength. It is impracticable to do this on distant signals because of fading. The most reliable method is to erect a temporary antenna of the type recommended for field-strength measurements (Fig. 4-66) and excite it from a low-power oscillator. The same precautions with respect to distance between the two antennas apply.

In this method, as in the one where the transmitting antenna is excited, it is necessary to minimize line radiation and pick-up if the results are to be reliable. The same tests may be applied. However, it is less easy to keep the s.w.r. under control. In the receiving case the s.w.r. on the transmission line depends on the load presented by the *receiver* to the line. Under most conditions the s.w.r. will be reasonably constant over an amateur band, although its value may not be known. However, the energy transfer from the antenna to the line depends on the mismatch between the driven element and the line. There is no convenient way to check this in the receiving case. About all that can be done is to apply power to the array after a set of tuning conditions has been reached, and then rematch at the driven element if necessary. After rematching the measurement will have to be repeated. Thus double checking is necessary if the results are to be comparable with those obtained by the field-strength method.

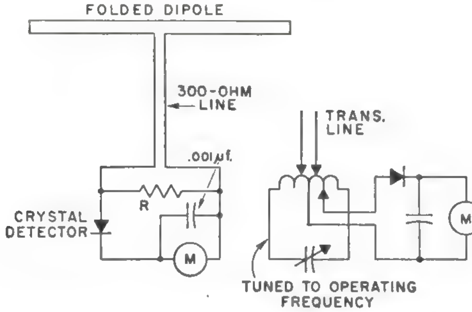


Fig. 4-66—Field-strength measurement set-up. The folded dipole should be at least as high as the antenna under test and should be three or more wave lengths away. *R* should be a 300-ohm carbon resistor to provide a proper load for the line, so that a line of any desired length can be used. If the sensitivity is not high enough with this arrangement the alternative connections at the right will result in increased meter readings. The taps are adjusted for maximum reading, keeping the transmission-line taps spaced equally on either side of the coil center-tap. The indicating meter, *M*, may be either a microammeter or 0–1 milliammeter.

THE QUAD ANTENNA

In this chapter it has been assumed that the various antenna arrays have been assemblies of linear half-wave (or approximately half-wave) dipole elements. However, other element forms may be used according to the same basic principles. For example, loops of various types may be combined into directive arrays. A popular type of parasitic array using loops is the "quad" antenna, in which loops having a perimeter of one wavelength are used in much the same way as dipole elements in the Yagi antenna.

The full-wave loop has been discussed in Chapter Two. Two such loops, one as a driven element and one as a reflector, are shown in Fig. 4-67. This is the original version of the quad; in subsequent development, loops tuned as directors have been added in front of the driven element. The square loops may be mounted either with the corners lying on horizontal and vertical lines, as shown at the left, or with two sides horizontal and two vertical (right). The feed points shown for these two cases will result in horizontal polarization, which is commonly used.

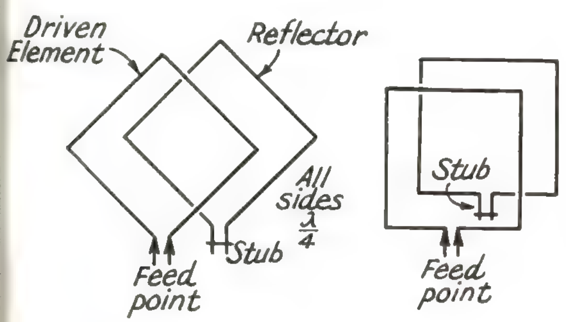


Fig. 4-67—The basic quad antenna, with driven loop and reflector loop. The loops are electrically one wavelength in circumference (¼ wavelength on a side). Both configurations shown give horizontal polarization; for vertical polarization, the driven element should be fed at one of the side corners in the arrangement at the left, or at the center of a vertical side in the "square" quad at the right.

The parasitic element is tuned in much the same way as the parasitic element in a Yagi antenna. That is, the parasitic loop is tuned to a lower frequency than the operating frequency when the element is to act as a reflector, and to a higher frequency when it acts as a director. Fig. 4-67 shows the parasitic element with an adjustable tuning stub, a convenient method of tuning since the resonant frequency can be changed simply by changing the position of the shorting bar on the stub. In practice, it has been found that the length around the loop should be approximately 3 percent greater than the self-resonant length if the element is a reflector, and about 3 percent shorter than the self-resonant length if the parasitic element is a director. Approximate formulas for the loop lengths in feet are

$$\textit{Driven element} = \frac{1005}{f_{\text{Mc.}}}$$

$$\textit{Reflector} = \frac{1030}{f_{\text{Mc.}}}$$

$$\textit{Director} = \frac{975}{f_{\text{Mc.}}}$$

for quad antennas intended for operation below 30 Mc. At v.h.f., where the ratio of loop circumference to conductor diameter is usually relatively small, the circumference must be increased in comparison to the wavelength. For example, a one-wavelength loop constructed of quarter-inch tubing for 144 Mc. should have a circumference about 4.5 percent greater than the wavelength in free space, as compared to the approximately 2 percent increase in the formula above for the driven element.

In any case, on-the-ground adjustment is required if optimum results are to be secured, especially with respect to front-to-back ratio. The method of adjustment parallels that outlined previously for the Yagi antenna.

Element spacings of the order of 0.14 to 0.2 wavelength are generally used, the smaller spacings being employed in antennas having more than two elements, where the structural support for elements with larger spacings tends to become difficult. The feed-point impedances of antennas having element spacings of this order have been found to be in the 40-to 60-ohm range, so the driven element can be fed through coaxial cable with only a small mismatch. For spacings of the order of 0.25 wavelength (physically feasible for two elements, or for several elements at 28 Mc.) the impedance more closely approximates the impedance of a driven loop alone (see Chapter Two)—that is, 80 or 90 ohms.

The feed methods described in Chapter Three can be used, just as in the case of the Yagi.

Directive Patterns and Gain

The small gain of a one-wavelength loop over a half-wave dipole (see Chapter Two) carries over into arrays of loops, that is, if a quad parasitic array and a Yagi having the same overall length (boom length) are compared the quad will have approximately 2 db. greater gain than the Yagi. This assumes that both antennas have the optimum number of elements for the antenna length; the number of elements is not necessarily the same in both when the antennas are long. Fig. 4-68 shows the results of comparative experimental measurements on the two antenna types at 440 Mc., made by J. E. Lindsay, Jr., WØHTH. The curves show

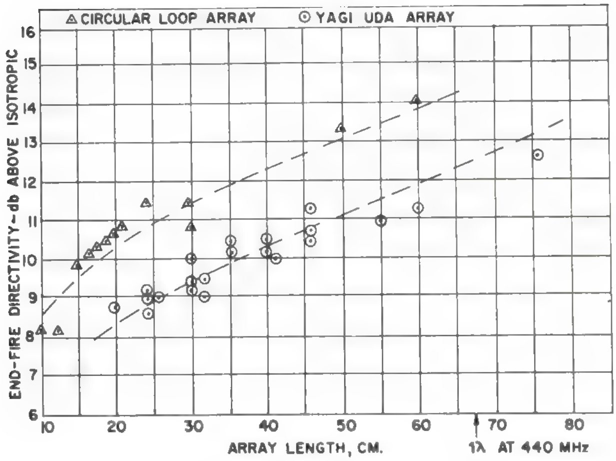


Fig. 4-68—Comparative directivity (and gain) of the Yagi and quad as a function of overall array length. Although measured with circular loops, the performance with the square loops used in the quad is comparable. The measurements were made on model antennas at 440 Mc. (WØHTH) but also apply at lower frequencies.

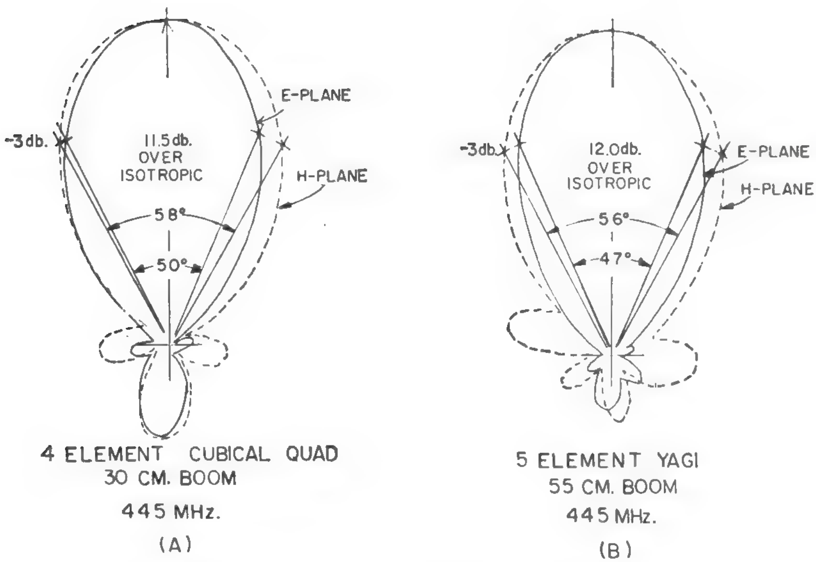


Fig. 4-69—Measured patterns of 4-element quad and 5-element Yagi antennas, showing approximately equivalent beamwidths. Measurements made on model antennas at 445 Mc. by WØHTH.

directivity vs. array length, which may be read as feet instead of centimeters if the frequency is taken as 14 Mc. instead of 440 Mc. Gain over a half-wave dipole (assuming negligible ohmic loss) can be found by subtracting 2.14 db. The gains were calculated from the measured patterns shown in Figs. 4-69 and 4-70.

The experimentally-measured patterns in Figs. 4-69 and 4-70 show that the beamwidths are approximately the same for the quad and

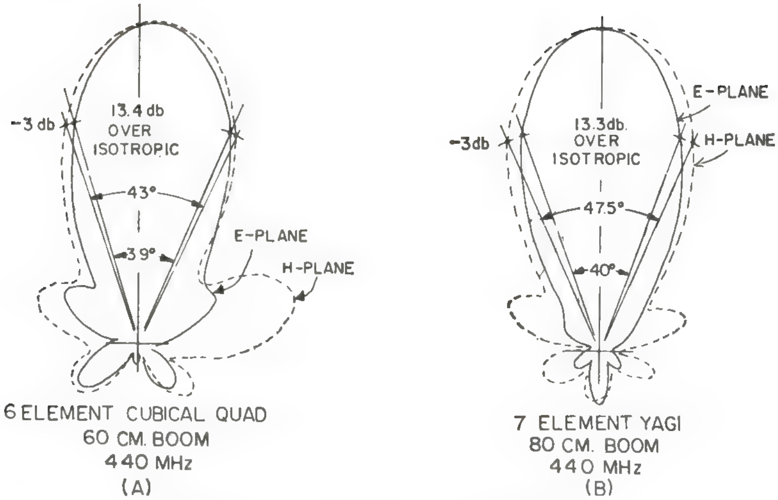


Fig. 4-70—Similar to Fig. 4-69, but for 6-element quad and 7-element Yagi antennas.

Yagi when the overall length of the latter is about twice (more closely, 1.8 times) the length of the quad, indicating that under such conditions the gains are about equal. Also, since the patterns are similar, both types will perform the same with respect to the vertical angle of radiation when mounted, with the same polarization, at the same height above ground.

Long-Wire Antennas

The power gain and directive characteristics of the harmonic wires (which are "long" in terms of wavelength) described in Chapter Two make them useful for long-distance transmission and reception on the higher frequencies. In addition, long wires can be combined to form antennas of various shapes that will increase the gain and directivity over a single wire. The term "long wire" as used in this chapter means any such configuration, not just a straight-wire antenna.

Long Wires vs. Multielement Arrays

In general, the gain achieved with long-wire antennas is not as great, when the space available for the antenna is limited, as can be secured from the multielement arrays in Chapter Four. To offset this the long-wire antenna has advantages of its own.

The construction of long-wire antennas is simple both electrically and mechanically, and there are no especially critical dimensions or adjustments. The long-wire antenna will work well and give satisfactory gain and directivity over a 2-to-1 frequency range; in addition, it will accept power and radiate it well on any frequency for which its over-all length is not less than about a half wavelength. Since a wire is not "long," even at 28 Mc., unless its length is at least equal to a half wavelength on 3.5 Mc., any long-wire antenna can be used on all amateur bands that are useful for long-distance communication.

As between two directive antennas having the same theoretical gain, one a multi-element array and the other a long-wire antenna, many amateurs have found that the long-wire antenna seems more effective in reception. One possible explanation is that there is a diversity effect with a long-wire antenna because it is spread out over a large distance rather than being concentrated in a small space; this may raise the average level of received energy for ionospheric propagation. Another factor is that long-wire antennas have directive patterns that are sharp in both the horizontal and vertical planes, and tend to concentrate the radiation at the low vertical angles that are most useful at the higher frequencies. This is not true of some types of multi-element arrays.

General Characteristics of Long-Wire Antennas

Whether the long-wire antenna is a single wire running in one direction or is formed into a V, rhombic, or some other configuration, there are certain general principles that apply and some performance features that are common to all types.

The first of these is that the power gain of a long-wire antenna as compared with a half-wave dipole is not considerable until the antenna is really "long"—long, that is, when the lengths are measured in wavelengths rather than in a specific number of feet. The reason for this is that the fields radiated by elementary lengths of wire along the antenna do not combine, at a distance, in as simple a fashion as the fields from half-wave dipoles used as described in Chapter Four. There is no point in space, for example, where the distant fields from all points along the wire are exactly in phase (as they are, in the optimum direction, in the case of two or more collinear or broadside dipoles when fed with in-phase currents). Consequently, the field strength at a distance always is less than would be obtained if the same length of wire were cut up into properly-phased separately-driven dipoles. As the wire is

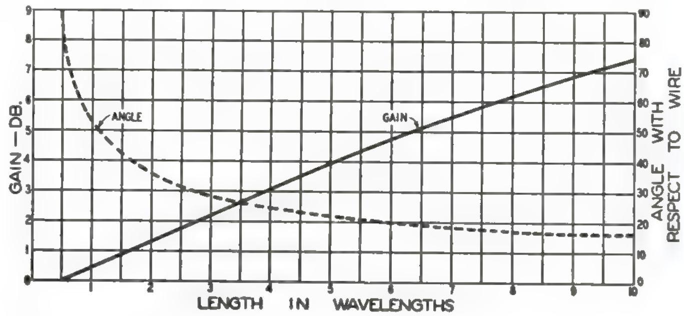


Fig. 5-1—Theoretical gain of a long-wire antenna over a dipole as a function of wire length. The angle, with respect to the wire, at which the radiation intensity is maximum also is shown.

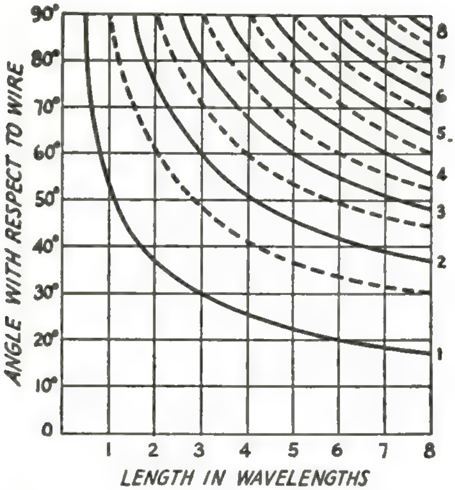


Fig. 5-2—Angles at which radiation from long wires is maximum (solid curves) and zero (broken curves). The major lobe, No. 1, has the power gains given by Fig. 5-1. Secondary lobes have smaller amplitude, but the maxima may exceed the radiation intensity from a half-wave dipole.

made longer the fields combine to form an increasingly intense main lobe, but this lobe does not develop rapidly until the wire is several wave lengths long. The longer the antenna, the sharper the lobe becomes; and since it is really a cone of radiation about the wire in free space, it becomes sharper in all planes. Also, the greater the length the smaller the angle with the wire at which the maximum radiation occurs.

Because many points along a long wire are carrying currents in different phase (usually with different current amplitude as well) the field pattern at a distance becomes more complex as the wire is made longer. This complexity is manifested in a series of minor lobes, the number of which increases with the wire length. The intentensity of radiation from the minor lobes is frequently as great as, and sometimes greater than, the radiation from a half-wave dipole. The energy radiated in the minor lobes is not available to improve the gain in the major lobe, which is another reason why a long-wire antenna has to be long to give appreciable gain.

Driven and parasitic arrays of the simple types described in Chapter Four do not have minor lobes of any great consequence. For that reason they frequently seem to have better directivity than long-wire antennas, because their response in directions other than that at which the antenna is aimed is well down. This will be so even if a multielement array and a long-wire antenna have the same actual gain in the favored direction. For amateur work, particularly with directive antennas that cannot be rotated, the minor lobes of a long-wire antenna have some advantages. In most directions the antenna will be as good as a half-wave dipole, and in addition will give high gain in the most favored direction; thus a long-wire antenna (depending on the design) frequently is a good all-around radiator in addition to being a good directive antenna.

In the discussion of directive patterns of long-wire antennas in this chapter, it should be kept in mind that the radiation patterns of resonant long wires are based on the assumption that each half-wave section of wire carries a current of the same amplitude. As pointed out in Chapter Two, this is not exactly true, since energy is radiated as it travels along the wire. For this reason it is to be anticipated that, although the theoretical pattern is bidirectional and identical in both directions, actually the radiation (and reception) will be best in one direction. This effect becomes more marked as the antenna is made longer.

Wave Angles

The wave angle at which maximum radiation takes place from a long-wire antenna depends on the same factors that operate in the case of simple dipoles and multielement antennas. That is, the directive pattern in the presence of ground is found by multiplying the free-space vertical-plane pattern of the antenna by the ground-reflection factors for the particular antenna height used. These factors are discussed in Chapter Two.

As mentioned a few paragraphs ago, the free-space radiation pattern of a long-wire antenna has a major lobe that forms a cone around the wire. The angle at which maximum radiation takes place becomes smaller, with respect to the wire, as the wire length is increased. For this reason a long-wire antenna is primarily a low-angle radiator when installed horizontally above the ground. Its performance in this respect is improved by selecting a height that also tends to concentrate the radiation at low wave angles.

Antenna systems formed from ordinary horizontal dipoles that are not stacked have, in most cases, a rather broad vertical pattern; the wave angle at which the radiation is maximum therefore depends chiefly on the antenna height. However, with a long-wire antenna the wave angle at which the major lobe is maximum can never be as great as the angle at which the first null occurs (see Fig. 5-2) even if the height is very low. (The efficiency may be less at very small heights, partly because of increased losses in the ground and partly because the pattern is affected in such a way as to put a greater proportion of the total power into the minor lobe.) The result is that when radiation at wave angles below 15 or 20 degrees is under consideration a long-wire antenna is less sensitive to height than are the multielement arrays or a simple dipole. To assure good results, however, the antenna should have a height equivalent to at least a half wavelength at 14 Mc.—that is, a minimum height of about 30 feet. Greater heights will give a worth-while improvement at wave angles below 10 degrees.

With an antenna of fixed physical length and

height, both length and height increase, in terms of wavelength, as the frequency is increased. The over-all effect is that both the antenna and the ground reflections tend to keep the system operating at high effectiveness throughout the frequency range. At low frequencies the wave angle is raised, and high wave angles are useful at 7 and 3.5 Mc. At high frequencies the converse is true. Good all-around performance usually results on all bands when the antenna is designed to be optimum in the 14-Mc. band.

Calculating Length

In this chapter lengths are always discussed in terms of wavelength. There can obviously be nothing very critical about wire lengths in an antenna system that will work over a frequency range including several amateur bands. The antenna characteristics change very slowly with length, except when the wires are short (i.e., around one wavelength) and there is no need to try to establish exact resonance at a particular frequency.

The formula for harmonic wires given in Chapter Two is quite satisfactory for determining the lengths of any of the antenna systems to be described. For convenience, the formula is given here in the following form:

$$\textit{Length} \text{ (feet)} = \frac{984\ (N - 0.025)}{\textit{Freq.} \text{ (Mc.)}}$$

where N is the number of full waves on the antenna. In cases where exact resonance is desired for some reason (for obtaining a resistive load for a transmission line at a particular frequency, for example) it is best established by trimming the wire length until measurement of the resonant frequency shows it to be correct.

Long Single Wires

The directional characteristics of wires that are multiples of a half wave in length have been discussed in Chapter Two. For convenience, the power gain over a dipole in free space and the angle with the wire at which the radiation is maximum are repeated in Fig. 5-1. The solid curve shows that the gain in decibels increases almost linearly with the length of the antenna. The gain does not become appreciable until the antenna is about four wavelengths long, where it is equivalent to doubling the transmitter power (3 db.). The actual gain over a half-wave dipole when the antenna is at a practical height above ground will depend on the way in which the radiation resistances of the long-wire antenna and the comparison dipole are affected by the height. This point was discussed in connection with the half-wave dipole in Chapter Two. The exact way in which the radiation resistance of a long wire varies with height depends on its length. In general, the resistance does not fluctuate as much, in terms of percentage, as does the resistance of a half-wave antenna. This is particularly true at heights from one-half wavelength up. Consequently, at a particular height the gain over a half-wave antenna at the same height will depend principally on how much the radiation resistance of the latter departs from the free-space value of 73 ohms on which the gain curve of Fig. 5-1 is based. The gain will be greatest at heights that make the dipole radiation resistance reach its highest values, approximately.

The minor lobes in the directive pattern have a maximum intensity approximately equivalent to that from a half-wave dipole. The nulls bounding the lobes are fairly sharp and are frequently somewhat obscured, in practice, by irregularities in the pattern. The locations of nulls and maxima for antennas up to eight wavelengths long are shown in Fig. 5-2.

Orientation

The broken curve of Fig. 5-1 shows the angle with the wire at which the radiation intensity is maximum. As shown in Chapter Two, there are two main lobes to the directive patterns of long-wire antennas; each makes the same angle with respect to the wire and each, considered in free space, is the solid formed by rotating the wire on its axis.

When the antenna is mounted horizontally above the ground, the situation depicted in Fig. 5-3 exists. Only one of the two lobes is considered in this drawing, and its lower half is cut off by the ground. The maximum intensity of radiation in the remaining half occurs through the broken-line semicircle; that is, the angle (B between the wire direction and the line marked "wave direction" is the angle given by Fig. 5-1 for the particular antenna length used.

In the practical case, there will be some wave angle (A) that is optimum for the frequency and the distance between the transmitter and receiver. Then for that wave angle the wire direction and the optimum geographical direction of transmission are related by the angle C. If the wave angle is very low, B and C will be practically equal. But as the wave angle becomes higher the angle C becomes smaller; in other words, the best direction of transmission and the direction of the wire more nearly coincide. They coincide exactly when C is zero; that is, when the wave angle is the same as the angle given by Fig. 5-1.

The maximum radiation from the antenna can be aligned with a particular geographical direction at a given wave angle by means of the following formula:

$$\cos C = \frac{\cos B}{\cos A}$$

In most amateur work the chief requirement is that the wave angle should be as low as possible, particularly at 14 Mc. and above. In such case it is usually satisfactory to make angle *C* the same as is given by Fig. 5-1.

It should be borne in mind that only the *maximum* point of the lobe is represented in Fig. 5-3. Radiation at higher or lower wave angles in any given direction will be proportional to the way in which the actual pattern shows the field strength to vary as compared with the maximum point of the lobe.

Tilted Wires

Fig. 5-3 shows that when the wave angle is equal to the angle which the maximum intensity of the lobe makes with the wire the best transmitting or receiving direction is that of the wire itself. If the wave angle is less than the lobe angle the best direction can be made to coincide with the direction of the wire by tilting the wire enough to make the lobe and wave angle coincide. This is shown in Fig. 5-4, for the case of a one-wave-length antenna tilted so that the maximum radiation from one lobe is horizontal to the left, and from the other is horizontal to the right (zero wave angle). The solid pattern can be visualized by imagining the plane diagram rotating about the antenna as an axis.

Since the antenna is neither vertical nor horizontal in this case, the radiation is part horizontally polarized and part vertically polarized. Computing the effect of the ground becomes complicated, because the horizontal and vertical components must be handled separately. In general, the directive pattern at any given wave angle becomes unsymmetrical when the antenna is tilted. For small amounts of tilt (less than the amount that directs the lobe angle horizontally) and for low wave angles the effect is to shift the optimum direction closer to the line of the antenna. This is true in the direction in which the antenna slopes downward. In the opposite direction the low-angle radiation is reduced.

Feeding Long Wires

It has been pointed out in Chapter Three (Fig. 3-44) that a harmonic antenna can be fed only

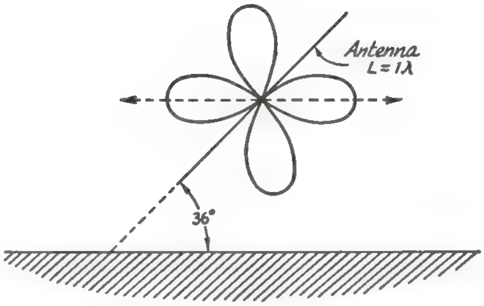


Fig. 5-4—Alignment of lobes for horizontal transmission by tilting a long wire in the vertical plane.

at the end or at a current loop. Since a current loop changes to a node when the antenna is operated at any even multiple of the frequency for which it is designed, a long-wire antenna will operate as a true long wire on all bands only when it is fed at the end.

A common method of feeding is to use a resonant open-wire line as shown at A in Fig. 5-5. This system will work on all bands down to the one, if any, at which the antenna is only a half wave long. Any convenient line length can be used if the transmitter is matched to the line input impedance by the methods described in Chapter Three.

Two arrangements for using non-resonant lines are given at B and C. The one at B is useful for one band only since the matching section must be a quarter wave long, approximately) unless a different matching section is used for each band. In C, the Q-section impedance should be adjusted to match the antenna to the line as described in Chapter Three, using the value of radiation resistance given in Fig. 2-23. This method is best suited to working with a 600-ohm transmission line. Although it will work as designed only on one band, the antenna can be used on other bands by treating the line and matching transformer as a resonant line. In such case, as mentioned earlier, the antenna will not radiate as a true long wire on even multiples of the frequency for which the matching system is designed.

The end-fed arrangement, although the most convenient when tuned feeders are used, suffers the disadvantage that there is likely to be a considerable antenna current on the line, as described in Chapter Three. In addition, the antenna reactance changes rapidly with frequency for the reasons outlined in Chapter Two (Figs. 2-9 and 2-10). Consequently, when the wire is several wavelengths long a relatively small change in frequency—a fraction of the width of a band—may require major changes in the adjustment of the transmitter-to-line coupling apparatus. Also, the line becomes unbalanced at all frequencies between those at which

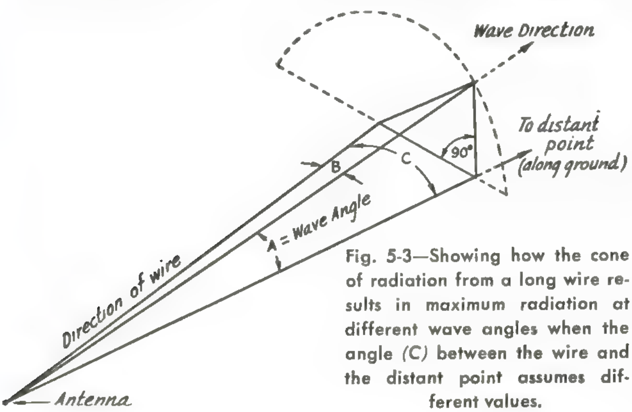


Fig. 5-3—Showing how the cone of radiation from a long wire results in maximum radiation at different wave angles when the angle (C) between the wire and the distant point assumes different values.

the antenna is exactly resonant. This leads to a considerable amount of radiation from the line. The unbalance can be overcome by using two wires in one of the arrangements described in succeeding sections.

Fig. 5-5—Methods of feeding long single-wire antennas.

Combinations of Resonant Long Wires

The directivity and gain of long wires may be increased by using two wires so placed in relation to each other as to make the fields from both combine to produce the greatest possible field strength at a distant point. The principle is similar to that used in forming the multielement arrays described in Chapter Four. However, the maximum radiation from a long wire occurs at an angle of less than 90 degrees with respect to the wire, so different physical relationships have to be used.

Parallel Wires

One possible method of using two (or more) long wires is to place them in parallel, with a spacing of ½ wavelength or so, and feed the two in phase. In the direction of the wires the fields will add up in phase. However, since the wave angle is greatest in the direction of the wire, as shown by Fig. 5-3, this method will result in rather high-angle radiation unless the wires are several wavelengths long. The wave angle can be lowered, for a given antenna length, by tilting the wires as described earlier. With a parallel arrangement of this sort the gain should be about 3 db. over a single wire of the same length, at spacings in the vicinity of ½ wavelength.

ECHELON ANTENNA

A better arrangement, and one that illustrates the principle of combining parallel long wires to obtain desired directive effects, is shown in Fig. 5-6. In this case the wires are still parallel but are also staggered; that is, the wires form two sides of a parallelogram but not of a rectangle. *A* and *B* are corresponding points on each of the wires. All the lines shown in Fig. 5-6 are in the same plane, that containing the antenna wires. The angle α between *YY* and the upper wire is made equal to the angle at which the radiation from a wire of that length is maximum, as given by Fig. 5-1. The angle between the wire and *XX* is the same, when *XX* lies in the direction of maximum radiation from the other pair of lobes in the plane directive diagram. *Y′Y′* is parallel to *XX*, since it is assumed that any point considered along *YY* will be so far distant that the waves from *A* and *B* follow parallel paths; similarly for any point along *XX*.

If the stagger between the wires is such that the line joining *A* and *B* is perpendicular to *YY*, waves radiated from *A* and *B* will travel exactly the same distance to reach any distant point along *YY*. This will be true of any pair of corresponding points along the antennas. If the two wires are fed in phase and carry the same currents, the fields at the distant point will add in phase. However, if the two wires are fed equal currents out of phase, the fields at a distant point

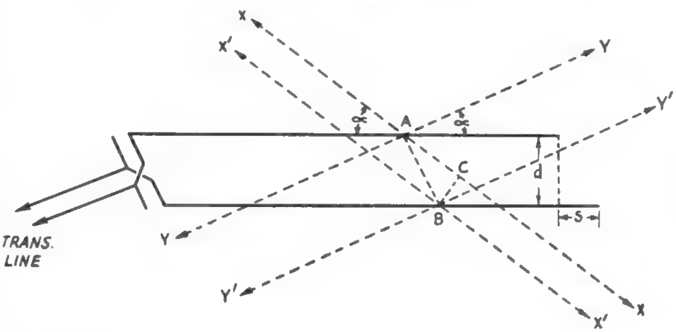


Fig. 5-6—The echelon antenna, using two parallel long wires. The effect of the stagger distance, *s*, and spacing, *d*, on the directive pattern are discussed in the text.

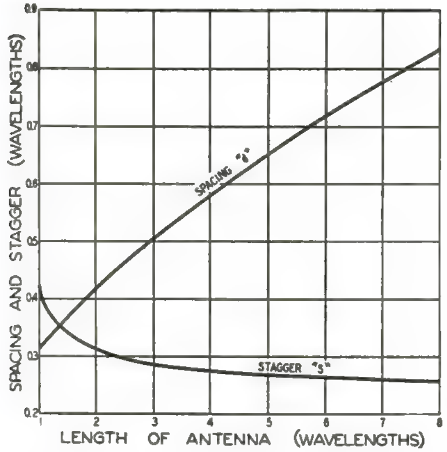


Fig. 5-7—Chart for determining stagger distance and spacing of an echelon antenna as a function of wire length.

along YY will cancel.

Now if the distance between the wires is such that the distance AC (BC being perpendicular to XX) is equal to one-half wavelength, there will be a difference of exactly one-half wavelength in the distance traveled by waves from A and B to reach a distant point along XX. Again this will be true of any pair of corresponding points along the wires. If the two wires are fed equal currents out of phase, the additional 180-degree phase delay because of the spacing makes the waves arrive in phase at the distant point. However, if the wires are fed equal currents in phase the delay caused by the spacing makes the waves cancel along the line XX.

Thus by using long wires in this "echelon" arrangement the radiation can be increased in one direction, say XX, and simultaneously canceled in the other, YY. By changing the phasing, YY will be the favored direction and cancellation will take place along XX. The preferred method is to feed the two wires out of phase, since this permits the transmission line to be balanced as shown in Fig. 5-6, and the mutual impedance between the wires is such as to increase the gain. In such case the maximum direction is XX and the lobes that would lie along YY, in the case of a single wire, will be eliminated. However, in a practical installation these directions can be interchanged, if desired, simply by changing the phasing of the elements. The principles discussed in Chapter Four in connection with element phasing apply here equally well.

The amount of stagger can be calculated from

$$s\ (\text{feet}) = \frac{492 \sin \alpha}{f\ (\text{Mc.}) \sin 2\alpha}$$

and the distance between wires is given by

$$d\ (\text{feet}) = \frac{492 \cos \alpha}{f\ (\text{Mc.}) \sin 2\alpha}$$

where α = angle of maximum radiation from a single wire. Fig. 5-7 gives the information in graphical form, with the lengths expressed in fractions of a wavelength.

Using wires in echelon has one disadvantage: the fact that the spacing and stagger distances are critical with respect to both wavelength and the length of the wires confines the operation to the frequency band for which the antenna is designed. This is probably one reason why it has not had much amateur use, despite the fact that it offers a choice of two bidirectional beams. The broad frequency characteristic that constitutes one of the principal advantages of the single long wire is retained in the other configurations described below.

THE V ANTENNA

Instead of using two long wires parallel to each other they may be placed in the form of a horizontal V, with the angle at the apex of the V twice the angle given by Fig. 5-1 for the particular length of wire used. The currents in the two wires should be out of phase. Under these conditions the plane directive patterns of the individual wires combine as is indicated in Fig. 5-8. Along a line in the plane of the antenna and bisecting the V the fields from the individual wires reinforce each other at a distant point. The other pair of lobes in the plane pattern is more or less eliminated, so that the pattern becomes essentially bidirectional.

The directional pattern of an antenna of this type is sharper in both the horizontal and vertical planes than the patterns of the individual wires composing it. Maximum radiation in both planes is along the line bisecting the V. There are minor lobes in both the horizontal and vertical patterns but if the legs are long in terms of wavelength the amplitude of the minor lobes is small. When the antenna is mounted horizontally above the ground (the usual method) the wave angle at which the radiation from the major lobe is maximum is determined by the height, but cannot exceed the angle values shown in Fig. 5-1 for the leg length used. Only the minor lobes give high-angle radiation.

The gain and directivity of a V depend on the length of the legs. Fig. 5-9 shows the gain and gives the maximum apex angle to use with legs up to 10 wavelengths long. The gains shown in this figure are approximate only; the solid curve is simply the gain of a single wire plus 3 db. for the effect of the second wire. The actual gain will be modified by the mutual impedance between the sides of the V, and data are available for only two cases, for leg lengths of 1 and 8 wavelengths. Although the broken gain curve in Fig. 5-9 merely joins the theoretical gains for these two cases, it is probably closer to the actual gain than the solid curve.

When the leg length is small there is some advantage in reducing the apex angle of the V because this changes the mutual impedance in such a way as to increase the gain of the antenna. The approximate change realized is given by the broken section in the curve marked "angle." For

example, the optimum apex angle in the case of a one-wavelength (each leg) *V* antenna is 90 degrees.

Lobe Alignment

It is possible to align the lobes from the individual wires with a particular wave angle by the method described in connection with Fig. 5-3. At very low wave angles the change in the apex angle is extremely small; for example, if the desired wave angle is 5 degrees the apex angles given in Fig. 5-9 will not be reduced more than a degree or so, even at the longest leg lengths which might be used.

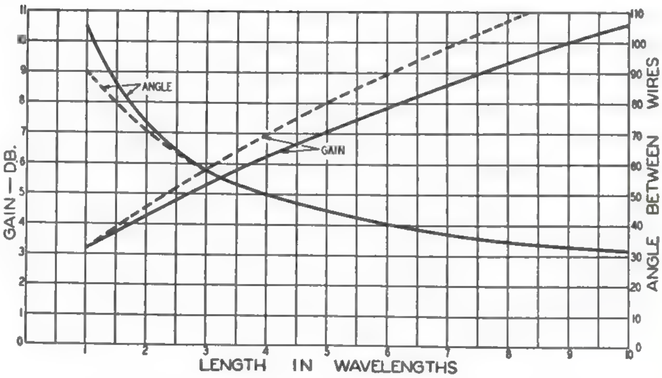


Fig. 5-9—Apex angle of V antenna as a function of the leg length in wavelengths. Curves of estimated gain also are shown. See text for discussion of these curves.

When the legs are long, alignment does not necessarily mean that the greatest signal strength will be secured at the wave angle for which the apex angle is chosen. It must be remembered that the polarization of the radiated field is the same as that of a plane containing the wire. As illustrated by the diagram of Fig. 5-3, at any wave angle other than zero the plane containing the wire and passing through the desired wave angle is not horizontal. In the limiting case where the wave angle and the angle of maximum radiation from the wire are the same the plane is vertical, and the radiation at that wave angle is vertically polarized. At in-between angles the polarization consists of both horizontal and vertical components.

When two wires are used in a *V* these planes have opposite slopes. In the plane bisecting the *V*, this makes the horizontally-polarized components of the two fields add together numerically, but the vertically-polarized components are out of phase and cancel completely. As the wave angle is increased the horizontally-polarized components become smaller, so the intensity of horizontally-polarized radiation decreases. On the other hand, the vertically-polarized components become more intense but always cancel each other. The over-all result is that although alignment for a given wave angle will increase the useful radiation at that angle, the wave angle at which maximum radiation occurs (in the direction of the line bisecting the *V*) is always below the wave angle for which the wires are aligned. As shown by Fig. 5-10, the difference between the apex angles required for optimum alignment of the lobes at wave angles of zero and 15 degrees is rather small, even when the legs are many wavelengths long.

For long-distance transmission and reception the lowest possible wave angle usually is the best. Consequently, it is good practice to choose an apex angle between the limits represented by the two curves in Fig. 5-10. The actual wave angle at which the radiation is maximum will depend on the shape of the vertical pattern and the height of the antenna above ground.

Multiband Design

When a *V* antenna is used over a range of frequencies—such as 14 to 28 Mc.—its characteristics over the frequency range will not change greatly if the legs are sufficiently long at the lowest frequency. The apex angle, at zero wave angle, for a 5-wavelength *V* (each leg approximately 350 feet long at 14 Mc.) is 44 degrees. At 21 Mc., where the legs are 7.5 wavelengths long, the optimum angle is 36 degrees, and at 28 Mc. where the *leg* length is 10 wavelengths it is 32 degrees. Such an antenna will operate well on all three frequencies if the apex angle is about 35 degrees. From Fig. 5-10, a 35-degree apex angle with

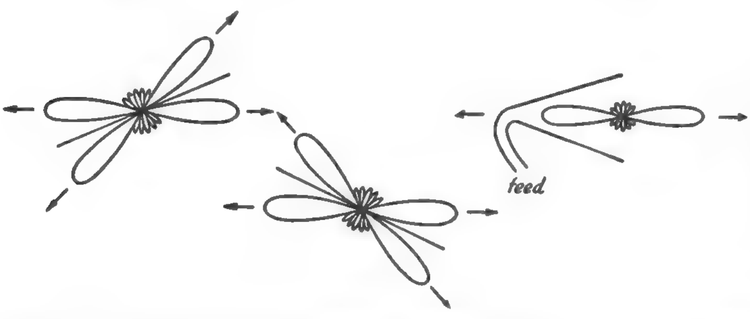


Fig. 5-8—Two long wires and their respective patterns are shown at the left. If these two wires are combined to form a V whose apex angle is twice that of the major lobes of the wires, and the wires are excited out of phase, the radiation along the bisector of the V adds and the radiation in the other directions tends to cancel.

a 5-wavelength *V* will align the lobes at a wave angle of something over 15 degrees, but this is not too high when it is kept in mind that the maximum radiation actually will be at a lower angle. At 28 Mc. the apex angle is a little large, but the chief effect will be a small reduction in gain and a slight broadening of the horizontal pattern, together with a tendency to reduce the wave angle at which the radiation is maximum. The same antenna can be used at 7 Mc. and 3.5 Mc., and on these bands the fact that the wave angle is raised is of less consequence, since high wave angles are useful. The gain will be small, however, because the legs are not long at these frequencies.

Other V Combinations

The gain can be increased about 3 db. by stacking two *Vs* one above the other, a half wavelength apart, and feeding them so that the legs on one side are in phase with each other and out of phase with the legs on the other side. This will result in a lowered angle of radiation. The bottom *V* should be at least a quarter wavelength above the ground and preferably a half wavelength.

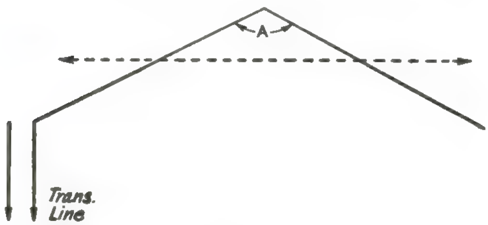


Fig. 5-11—The obtuse-angle V antenna. Angle A is equal to 180 degrees minus the angle given in Fig. 5-9.

Two *V* antennas can be broadsided to form a "W," giving an additional 3-db. gain. However, two transmission lines are required and this, plus the fact that five poles are needed to support the system, renders it normally impractical for the amateur.

The *V* antenna can be made unidirectional by using another *V* placed an odd multiple of a quarter wavelength in back of the first and exciting the two with a phase difference of 90 degrees. The system will be unidirectional in the direction of the antenna with the lagging current. However, the V reflector *is* not normally employed by amateurs at low frequencies because it restricts the use to one band and requires a fairly elaborate supporting structure. Stacked Vs with driven reflectors could, however, be built for the 200- to 500-Mc. region without much difficulty. The over-all gain for such an antenna (two stacked Vs, each with a *V* reflector) is about 6 db. greater than the gains given by the curves.

Feeding the V

The *V* antenna is most conveniently fed by tuned feeders, since they permit multiband operation. Although the length of the wires in a *V* beam is not at all critical, it is important that both wires be of the same electrical length. Balanced feeder currents (in a tuned line) give sufficient indication of balanced lengths in the antenna proper.

If it is desired to use a nonresonant line, probably the most appropriate matching system is that using a stub or quarter-wave matching section. The adjustment is as described in Chapter Three.

OBTUSE-ANGLE Vs

In the type of V just described the angle between the wires always is 90 degrees or less when the wires are more than 1 wavelength long. Another type of V can be formed by folding one wire as shown in Fig. 5-11 to align the lobes of maximum radiation. The angle *A* in this case is equal to 180 degrees less twice the angle given by Fig. 5-1 for a single wire having a length equal to that of one leg. For example, if each leg of the obtuse-angle *V* of Fig. 5-11 is 3-wavelengths (total length 6 wavelengths) the angle given by Fig. 5-1 is 29 degrees. The angle *A* in Fig. 5-11 will then be 180—58 = 122 degrees.

The obtuse angle *V* is seldom used in the form shown in Fig. 5-11 because twice the over-all length is required and the gain is less than with an acute-angle V having the same leg length. This is because of the

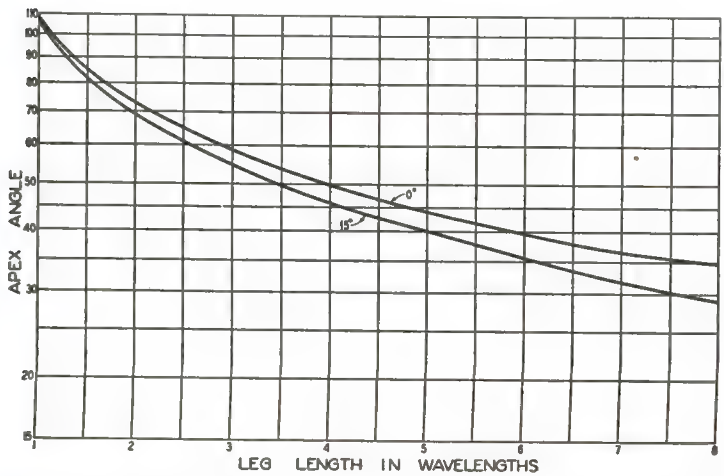


Fig. 5-10—Apex angle of V antenna for alignment of main lobe at different wave angles, as a function of leg length in wavelengths.

way the mutual impedance between legs compares in the two cases. However, the obtuse-angle V has the advantage that a change in frequency causes the major lobe from one leg to shift in one direction while the lobe from the other leg shifts the opposite way. This tends to make the optimum direction stay the same over a considerable frequency range. The pattern broadens and the gain is reduced when the antenna is operated at frequencies far removed from that for which it is designed.

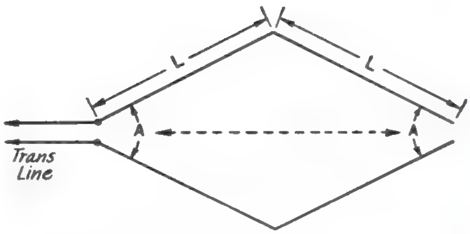


Fig. 5-12—The resonant rhombic or diamond-shaped antenna. All legs are the same length, and opposite angles of the diamond are equal.

THE RESONANT RHOMBIC ANTENNA

The diamond-shaped or rhombic antenna shown in Fig. 5-12 can be looked upon as two acute-angle Vs placed end-to-end or as two obtuse-angle Vs placed side-by-side. This arrangement has two advantages over the simple V that have caused it to be favored by amateurs. For the same total wire length it gives somewhat greater gain than the V; a rhombic 4 wavelengths on a leg, for example. And, because of the compensating effect mentioned in connection with the obtuse-angle V, the directional pattern is less affected by frequency when the antenna is used over wide frequency range. The disadvantage of the rhombic as compared with the V is that one additional support is required.

The same factors that govern the design of the V antenna apply in the case of the resonant rhombic. The angle *A* in the drawing is the same as that for a V (Fig. 5-9) having a leg length equal to *L*. If it is desired to align the lobes from individual wires with the wave angle, the curves of Fig. 5-10 may be used, again using the length of one leg in taking the data from the curves. The diamond-shaped antenna also can be operated as a nonresonant antenna, as described later in this chapter, and much of the discussion in that section applies to the resonant rhombic as well.

The direction of maximum radiation with a resonant rhombic is given by the arrows in Fig. 5-12; i.e., the antenna is bidirectional. There are minor lobes in other directions, their number and intensity depending on the leg length. When used at frequencies below the v.h.f. region the rhombic antenna is always mounted with the plane containing the wires horizontal. The polarization in this plane, and also in the perpendicular plane that bisects the rhombic, is horizontal. At 144 Mc. and above, the dimensions are such that the antenna can be mounted with the plane containing the wires vertical if vertical polarization is desired.

When the rhombic antenna is to be used on several amateur bands it is advisable to choose the apex angle, *A*, on the basis of the leg length in wavelengths at 14 Mc. This point is covered in more detail in connection both with the V and with the nonresonant rhombic. Although the gain on higher frequency bands will not be quite as favorable as if the antenna had been designed for the higher frequencies, the system will radiate well at the low angles that are necessary at such frequencies. At frequencies below the design frequency the greater apex angle of the rhombic (as compared with a V of the same total length) is more favorable to good radiation than in the case of the V.

The resonant rhombic antenna can be fed in the same way as the V. Resonant feeders are necessary if the antenna is to be used in several amateur bands.

Nonresonant Long-Wire Antennas

All the antenna systems previously considered have been based on resonant operation; that is, with standing waves of current and voltage along the wire. Although most antenna designs are based on using resonant wires, resonance is by no means a necessary condition for the wire to radiate and intercept electromagnetic waves.

In Fig. 5-13, let us suppose that the wire is parallel with the ground (horizontal) and is terminated by a load *Z* equal to its characteristic impedance, Z_0. The load *Z* can represent a receiver matched to the line. The resistor *R* is also equal to the Z_0 of the wire. A wave coming from the direction *Z* will strike the wire first at its far end and sweep across the wire at an angle until it reaches the end at which *Z* is connected .In doing so it will induce voltages in the antenna and currents will flow as a result. The current flowing toward *Z* is the useful output of the antenna, while the current flowing toward *R* will be absorbed in *R*. The same thing is true of a wave coming from the direction *X′*.

The greatest possible power will be delivered to the load *Z* when the individual currents induced as the wave sweeps across the wire all combine properly on reaching the load. The currents will reach *Z* in optimum phase when the time required for a current to flow from the far end of the antenna to *Z* is exactly one-half cycle longer than the time taken by the wave to sweep over the antenna. Since a half cycle is equivalent to a half wavelength in space, the length of the antenna must be one-half wavelength greater than the distance traversed by the wave

from the instant it strikes the far end of the antenna to the instant that it reaches the near end. This is shown by the small drawing, where *AC* represents the antenna, *BC* is a line perpendicular to the wave direction, and *AB* is the distance traveled by the wave in sweeping past *AC*. *AB* must be one-half wavelength shorter than *AC*. Similarly, *AB′* must be the same length as *AB* for a wave arriving from *X′*.

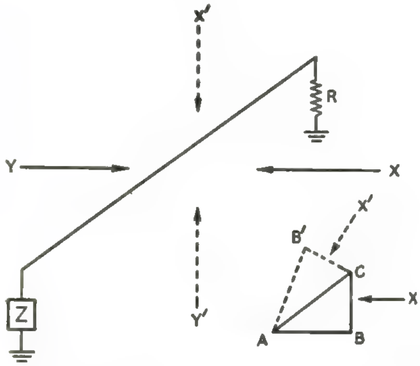


Fig. 5-13—Nonresonant long-wire antenna.

A wave arriving at the antenna from the opposite direction *Y* (or *Y′*), will similarly result in the largest possible current at the *far* end. However, the far end is terminated in *R*, which is equal to Z, so all the power delivered to *R* by a wave arriving from *Y* will be absorbed in *R*. The current traveling to Z will produce a signal in Z in proportion to its amplitude. If the antenna length is such that all the individual currents arrive at Z in such phase as to add up to zero, there will be no current through Z. At other lengths the resultant current may reach appreciable values. The lengths that give zero amplitude are those which are odd multiples of ¼ wavelength, beginning at ¾ wavelength. The response from the *Y* direction is greatest when the antenna is any even multiple of ½ wavelength long; the higher the multiple, the smaller the response.

Directional Characteristics

The explanation above considered the phase but not the relative amplitudes of the individual currents reaching the load. When the appropriate correction is made, the angle with the wire at which radiation or response is maximum is given by the curve of Fig. 5-14. The response drops off gradually on either side of the maximum point, resulting in lobes in the directive pattern much like those for harmonic antennas, except that the system is substantially unidirectional. Typical patterns are shown in Fig. 5-15. When the antenna length is 3/2 wavelength or greater there are also angles at which secondary maxima occur; these secondary maxima (minor lobes) have their peaks approximately at angles for which the length *AB*, Fig. 5-13, is less than *AC* by any odd multiple of one-half wavelength. When *AB* is shorter than *AC* by an *even* multiple of a half wavelength the induced currents cancel each other completely at *Z*, and in such cases there is a null for waves arriving in the direction perpendicular to *BC*.

The antenna of Fig. 5-13 responds to horizontally-polarized signals when mounted horizontally. If the wire lies in a plane that is vertical with respect to the earth it responds to vertically-polarized signals. By reciprocity, the characteristics for transmitting are the same as for receiving. For average conductor diameters and heights above ground, the Z_0 of the antenna is of the order of 500 to 600 ohms.

It is apparent that an antenna operating in this way has much the same characteristics as a transmission line. When it is properly terminated at both ends there are traveling waves, but no standing waves, on the wire. Consequently the current is substantially the same all along the wire. Actually, it decreases in the direction in which the current is flowing because of energy loss by radiation as well as by ohmic loss in the wire and the ground. The antenna can be looked upon as a transmission line terminated in

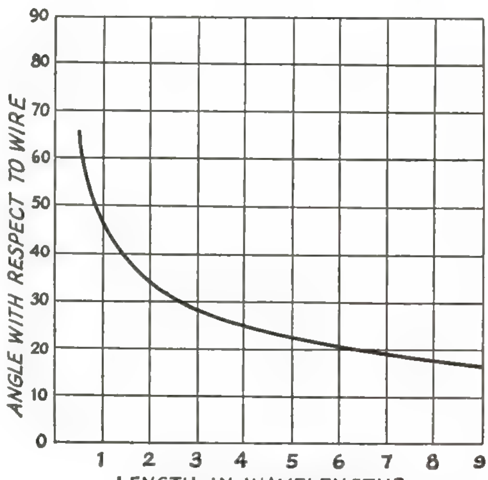


Fig. 5-14—Angle with respect to wire axis at which the radiation from a nonresonant long-wire antenna is maximum.

its characteristic impedance, but having such wide spacing between conductors (the second conductor in this case is the image of the antenna in the ground) that radiation losses are by no means inconsequential.

A wire terminated in its characteristic impedance will work on any frequency, but its directional characteristics change with frequency as shown by Fig. 5-14. To give any appre-

ciable gain over a dipole the wire must be at least a few wavelengths long. The angle at which maximum response is secured can be in any plane that contains the wire axis, so in free space the major lobe will be a cone. In the presence of ground, the discussion given in connection with Fig. 5-3 applies, with the modification that the angles of best radiation or response are those given in Fig. 5-14 rather than by Fig. 5-1. As comparison of the two curves will show, the difference in optimum angle between resonant and nonresonant wires is quite small.

THE INVERTED V ANTENNA

Two nonresonant long wires may be combined in much the same fashion as the resonant wires discussed earlier in this chapter. One example is the "inverted V" or "half rhombic" antenna shown in Fig. 5-16. For example, suppose that each of the sides, or legs, of the V has a length such that the major lobe of *each* wire makes an angle of 36 degrees with the wire. Then the "tilt angle." ϕ (half the total angle included between the wires), may be adjusted so that the major lobes of radiation from the two wires will add in the desired direction. The tilt angle then becomes the complement of the angle which the main lobe makes with each wire. This is shown in Fig. 5-16, for horizontal transmission to the right in the figure. In directions other than to the right in the plane containing the antenna the radiation from the individual wires will tend to cancel more or less completely.

For signals coming from the optimum direction the polarization is vertical when the plane containing the wires is vertical.

The advantage of combining wires as shown in Fig. 5-16 is the same as previously mentioned for the obtuse-angle V. That is, as the frequency is changed the optimum direction for one leg shifts in the opposite direction to that for the other leg. This tends to maintain the same direction of maximum response (or transmission) over a wide frequency range, although it is accompanied by a broadening of the pattern.

The form of tilted wire antenna shown in Fig. 5-16 is simple in construction, since only one pole is required. However, for all but quite high frequencies—28 Mc. and higher—the pole height needed to provide the proper tilt angle for a leg length great enough to provide appreciable gain is beyond the facilities available

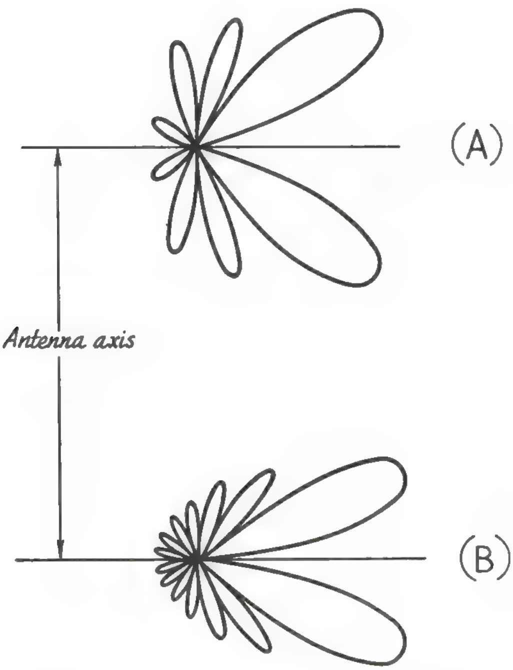


Fig. 5-15—Typical radiation patterns (cross section of solid figure) for terminated long wires. (A) length two wavelengths; (B) four wavelengths; both for an idealized case in which there is no decrease of current along the wire. In practice, the pattern is somewhat distorted by wire attenuation.

to most amateurs. Also, it is difficult to secure a satisfactory termination because of variation in ground resistance with weather conditions. This might be overcome by the use of a large ground screen under the antenna and extending a half wavelength or so beyond the wire in all directions, but the installation of such a screen probably would be impracticable in most locations. A form of transmission-line termination also may be used, with the far end of the antenna connected through the terminating resistor to the center of a half-wave wire running parallel to the ground and perpendicular to the line of the antenna. The impedance at the center of such a wire will be low, and currents induced by the incoming waves from the desired direction will balance out. However, such a termination is good only for the frequency for which the wire is cut, so that a wide frequency range is not possible. These difficulties are overcome by the use of the diamond-shaped, or rhombic, antenna.

THE NONRESONANT RHOMBIC ANTENNA

The nonresonant rhombic antenna, shown schematically in Fig. 5-17, consists of two tilted-wire antennas of the type shown in Fig. 5-16 placed side by side. The terminating resistor is connected between the far ends of the two sides, and is made approximately equal to the characteristic impedance of the antenna as a unit. The rhombic may be constructed either horizontally or vertically, but practically always is horizontal at frequencies below 54 Mc., since the pole height required is considerably less. Also, horizontal polarization is equally, if not more, satisfactory at these frequencies.

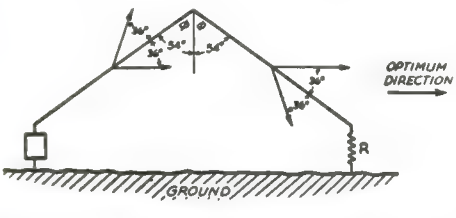


Fig. 5-16—The tilted-wire antenna, showing the directions of the main lobes for each wire for legs two wave lengths long, when the tilt angle ϕ is adjusted for alignment of the lobes.

The basic principle of combining lobes of maximum radiation from the four individual wires constituting the rhombus or diamond is the same in either the nonresonant type shown in Fig. 5-17 or the resonant type described earlier in this chapter. The included angles should differ slightly because of the differences between resonant and nonresonant wires, but comparison of Figs. 5-1 and 5-14 show that the differences are almost negligible.

Tilt Angle

It is a matter of custom, in dealing with the nonresonant or "terminated" rhombic, to talk about the "tilt angle" (ϕ in Fig. 5-17) rather than the angle of maximum radiation with respect to an individual wire. The tilt angle is simply 90 degrees minus the angle of maximum radiation. In the case of a rhombic antenna designed for zero wave angle the tilt angle is 90 degrees minus the values given in Fig. 5-14.

Fig. 5-18 shows the tilt angle as a function of the antenna leg length. The curve marked "O°" is for a wave angle of zero degrees; that is,

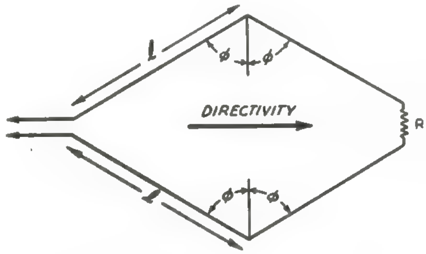


Fig. 5-17—The nonresonant rhombic antenna.

maximum radiation in the plane of the antenna. The other curves show the proper tilt angles to use when aligning the major lobe with a desired wave angle. For a wave angle of 5 degrees the difference in tilt angle is less than one degree for the range of lengths shown. Just as in the case of the resonant V and resonant rhombic, alignment of the wave angle and lobes always results in still greater radiation at a lower wave angle, and for the same reason, but also results in the greatest possible radiation at the desired wave angle.

The broken curve marked "Optimum Length" shows the leg length at which maximum gain is secured at a chosen wave angle. Increasing the leg length beyond the optimum will result in lessened gain, and for that reason the curves do not extend beyond the optimum length. Note that the optimum length becomes greater as the desired wave angle is smaller. Leg lengths over 6 wavelengths are not recommended because the directive pattern becomes so sharp that the antenna performance is highly variable with small changes in the angle, both horizontal and vertical, at which an incoming wave reaches the antenna. Since these angles vary to some extent in ionospheric propagation, it does not pay to attempt to use too great a degree of directivity.

Multiband Design

When a rhombic antenna is to be used over a considerable frequency range it is worth paying some attention to the effect of the tilt angle on the gain and directive pattern at various frequencies. For example, suppose the antenna is

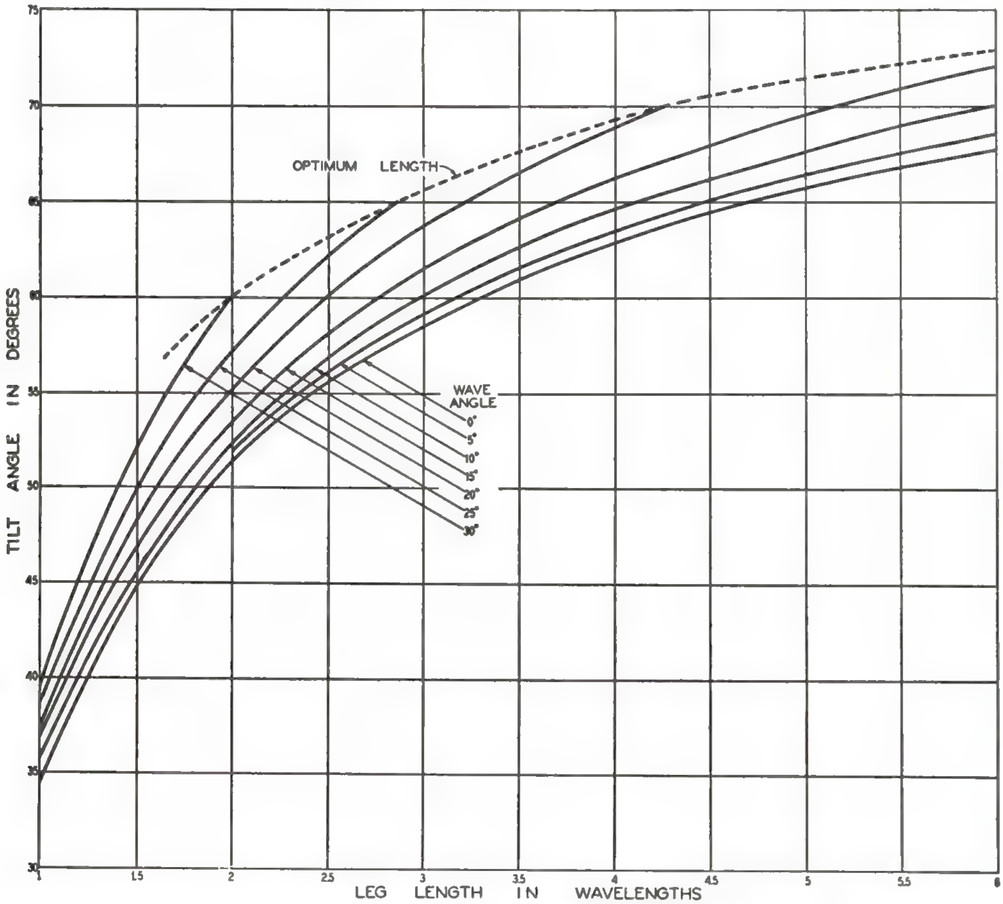


Fig. 5-18—Rhombic-antenna design chart. For any given leg length, the curves show the proper tilt angle to give maximum radiation at the selected wave angle. The broken curve marked "Optimum Length" shows the leg length that gives the maximum possible output at the selected wave angle. The optimum length as given by the curves should be multiplied by 0.74 to obtain the leg length for which the wave angle and main lobe are aligned.

to be used at frequencies up to and including the 28-Mc. band, and that the leg length is to be 6 wavelengths on the latter frequency. For zero wave angle the optimum tilt angle is 68 degrees, and the calculated free-space directive pattern in the vertical plane bisecting the antenna is shown in the right-hand drawing of Fig. 5-19. At 14 Mc. this same antenna has a leg length of three wavelengths, which calls for a tilt angle of 58.5 degrees for maximum radiation at zero wave angle. The calculated patterns for tilt angles of 58.5 and 68 degrees are shown in the left-hand drawing in Fig. 5-19, and it is seen that if the optimum tilt for 28-Mc. operation is used the gain will be reduced and the wave angle raised at 14 Mc. In an attempt at a compromise, we might select a wave angle of 15 degrees, rather than zero, for 14 Mc. since, as shown by Fig. 5-18, the tilt angle is larger and thus more nearly coincides with the tilt angle for zero wave angle on 28 Mc. From the chart, the tilt angle for 3 wavelengths on a leg and a 15-degree wave angle is 61.5 degrees. The patterns with this tilt angle are shown in Fig. 5-19 for both the 14- and 28-Mc. cases. The effect at 28 Mc. is to decrease the gain at zero wave angle by more than 6 db. and to split the radiation in the vertical plane into two lobes, one of which is at a wave angle too high to be useful at this frequency.

Inasmuch as the gain increases with the leg length in wavelengths, it is probably better to favor the lower frequency in choosing the tilt angle. In the present example, the best compromise probably would be to split the difference between the optimum tilt angle for the 15-degree wave angle at 14 Mc. and that for zero wave angle at 28 Mc.; that is, use a tilt angle of about 64 degrees.

The patterns of Fig. 5-19 are in the vertical plane through the center of the antenna only. In vertical planes making an angle with the

antenna axis the patterns may differ considerably. The effect of a tilt angle that is smaller than the optimum is to broaden the horizontal pattern, so at 28 Mc. the antenna in the example would be less directive in the horizontal plane than would be the case if it were designed for optimum performance at that frequency. It must also be pointed out that the patterns given in Fig. 5-19 are free-space patterns and must be multiplied by the ground-reflection factors for the actual antenna height used, if the actual vertical patterns is to be determined.

Power Gain

The theoretical power gain of a nonresonant rhombic antenna over a dipole, both in free space, is given by the curve of Fig. 5-20. This curve is for zero wave angle and includes an allowance of 3 db. for power dissipated in the terminating resistor. The actual gain of an antenna mounted horizontally above the ground, as compared with a dipole at the same height, can be expected to vary a bit either way from the figures given by the curve. The power lost in the terminating resistor is probably less than 3 db. in the average installation, since more than half of the power input is radiated before the end of the antenna is reached. However, there is also more power loss in the wire and in the ground under the antenna than in the case of a simple dipole, so the 3 db. figure is probably a representative estimate of *over-all* loss.

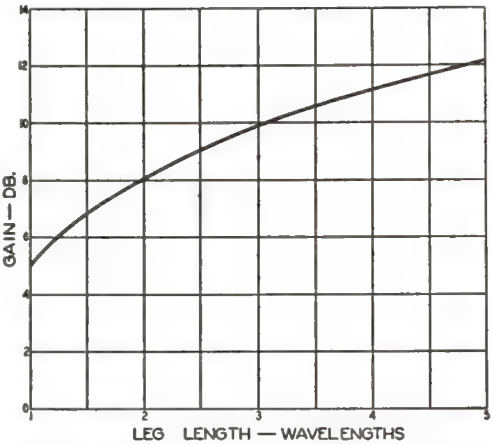


Fig. 5-20—Theoretical gain of a nonresonant rhombic antenna over a half-wave dipole in free space. This curve includes an allowance of 3 db. for loss in the terminating resistor.

Termination

Although there is no marked difference in the gain obtainable with resonant and nonresonant rhombics of comparable design, the nonresonant antenna has the advantage that over a wide frequency range it presents an essentially resistive and constant load to the transmitter coupling apparatus. In addition, nonresonant operation makes the antenna substantially unidirectional, while the "unterminated" or resonant rhombic is always bidirectional, although not symmetically so. In a sense, it can be considered that the power dissipated in the terminating resistor is simply power that would have been radiated in the other direction had the resistor not been there, so the fact that some of the power (about one-third) is used up in heating the resistor does not mean an actual loss in the desired direction.

The characteristic impedance of an ordinary rhombic antenna, looking into the input end, is of the order of 700 to 800 ohms when properly terminated in a resistance at the far end. The terminating resistance required to bring about the matching condition usually is slightly higher than the input impedance because of the loss of energy through radiation by the time the far end is reached. The correct value usually will be found to be of the order of 800 ohms, and should be determined experimentally if the flattest possible antenna is desired. However, for average work a noninductive resistance of 800 ohms can be used with the assurance that the operation will not be far from optimum.

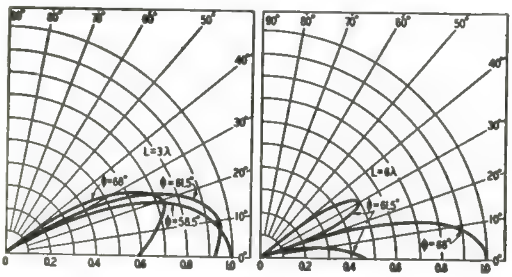


Fig. 5-19—Showing the effect of tilt angle on the free-space vertical pattern of a nonresonant rhombic antenna having a leg length of 3 wavelengths at one frequency and 6 wavelengths at twice the frequency. These patterns apply only in the direction of the antenna axis.

The terminating resistor must be practically a pure resistance at the operating frequencies; that is, its inductance and capacitance should be negligible. Ordinary wire-wound resistors are not suitable because they have far too much inductance and distributed capacitance. Small carbon resistors have satisfactory electrical characteristics but will not dissipate more than a few watts and so cannot be used, except when the transmitter power does not exceed 10 or 20 watts or when the antenna is to be used for reception only. The special resistors designed either for use as "dummy" antennas or for terminating rhombic antennas should be used in other cases. To allow a factor of safety, the total rated power dissipation of the resistor or resistors should be equal to half the power output of the transmitter.

To reduce the effects of stray capacitance it is desirable to use several units, say three, in series even when one alone will safely dissipate the power. The two outer units should be identical and each should have one-fourth to one-third the total resistance, with the center unit making up the difference. The units should be installed in a weatherproof housing at the end of the antenna to protect them and to permit mounting without mechanical strain. The connecting leads should be short so that little extraneous inductance is introduced.

Alternatively, the terminating resistance may be placed at the end of an 800-ohm line connected to the end of the antenna. This will permit placing the resistors and their housing at a point convenient for adjustment rather than at the top of the pole. The line length is not critical, since it operates without standing waves and hence is nonresonant.

Multiwire Rhombics

The input impedance of a rhombic antenna constructed as in Fig. 5-17 is not quite constant as the frequency is varied. This is because the varying separation between the wires causes the characteristic impedance of the antenna to vary along its length. The variation in Z_0 can be minimized by a conductor arrangement that increases the capacitance per unit length in proportion to the separation between the wires.

The method of accomplishing this is shown in Fig. 5-21. Three conductors are used, joined together at the ends but with increasing separation as the junction between legs is approached.

Fig. 5-21—Three-wire rhombic antenna. Use of multiple wires improves the impedance characteristic of a nonresonant rhombic.

As used in commercial installations having legs several wavelengths long, the spacing between wires at the center is 3 to 4 feet. Since all three wires should have the same length, the top and bottom wires will be slightly farther from the support than the middle wire. Using three wires in this way reduces the Z_0 of the antenna to approximately 600 ohms, thus providing a better match for a practicable open-wire line, in addition to smoothing out the impedance variations over the frequency range.

A similar effect, although not quite so good, is obtained by using two wires instead of three. It has been found that, with the 3-wire system, the gain of the antenna is slightly greater (of the order of 1 db.) than when only a single conductor is used.

Front-to-Back Ratio

It is theoretically possible to obtain an infinite front-to-back ratio with a terminated rhombic antenna, and in practice very large values can actually be secured. However, when the antenna is terminated in its characteristic impedance the infinite front-to-back ratio can be secured only at frequencies for which the leg length is an odd multiple of a quarter wavelength, as described in the section on nonresonant long wires. The front-to-back ratio is smallest at frequencies for which the leg length is a multiple of a half wavelength.

When the leg length is not an odd multiple of a quarter wave at the frequency under consideration, the front-to-back ratio can be made very high by slightly decreasing the value of terminating resistance. This permits a small reflection from the far end of the antenna which cancels out, at the input end, the residual response. With large antennas the front-to-back ratio may be made very large over the whole frequency range by experimental adjustment of the terminating resistance. Modification of the terminating resistance also permits "steering" the back null over a small horizontal range so that signals coming from a particular spot not exactly to the rear of the antenna may be minimized.

Ground Effects

Reflections from the ground play exactly the same part in determining the vertical directive pattern of a horizontal rhombic antenna that they play with other horizontal antennas. Consequently, if a low wave angle is desired it is necessary to make the height great enough to bring the reflection factor into the higher range of values given by the charts in Chapter Two. At 14 Mc. heights of ¾ to 1 wavelength are desirable. The antenna will work at heights as low as ½ wavelength, of course, but with reduced radiation at the lower angles, particularly below 10 degrees. In this respect it is under no greater or lesser handicap than any other antenna system that is close to the ground.

Alignment of Lobes, Wave Angle, and Ground Reflections

When maximum antenna response is desired at a particular wave angle (or maximum radiation is desired at that angle) the major lobe of the antenna cannot only be aligned with the wave angle as previously described but also with a maximum in the ground-reflection factor. When this is done it is no longer possible to consider the antenna height independently of other aspects of rhombic design. The wave

angle, leg length, and height become mutually dependent.

This method of design is of particular value when the antenna is built to be used over fixed transmission distances for which the optimum wave angle is known. It has had wide application in commercial work with nonresonant rhombic antennas, but seems less desirable for amateur use where, for the long-distance work for which rhombic antennas are built, the lowest wave angle that can be achieved is the most desirable. Alignment of all three factors is limited in application because it leads to impracticable heights and leg lengths for small wave angles. Consequently, when a fairly broad range of low wave angles is the objective it is more satisfactory to design for a low wave angle and simply make the antenna as high as possible.

Fig. 5-22 shows the lowest height at which ground reflections make the radiation maximum at a desired wave angle. It can be used in conjunction with Fig. 5-18 for complete alignment of the antenna. For example, if the desired wave angle is 20 degrees, Fig. 5-22 shows that the height must be 0.75 wavelength. From Fig. 5-18, the optimum leg length is 4.2 wavelengths and the tilt angle is just under 70 degrees. A rhombic antenna so designed will have the maximum possible output that can be obtained at a wave angle of 20 degrees; no other set of dimensions will be as good. However, it will have still greater output at some angle lower than 20 degrees, for the reasons given earlier. When it is desired to make the maximum output of the antenna occur at the 20-degree wave angle, it may be accomplished by using the same height and tilt angle, but with the leg length reduced by 26 per cent. Thus for such alignment the leg length should be $4.2 \times 0.74 = 3.1$ wavelengths. However, the output at the 20-degree wave angle will be smaller than with 4.2-wavelength legs, despite the fact that the smaller antenna has its maximum radiation at 20 degrees. The reduction in gain is about 1.5 db.

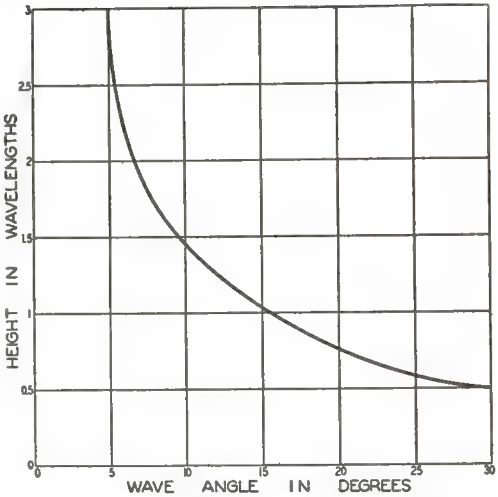


Fig. 5-22—Antenna height to be used for securing maximum radiation at a desired wave angle. This curve applies to any type of horizontal antenna.

Methods of Feed

If the broad frequency characteristic of the rhombic antenna is to be fully utilized the feeder system used with it must be similarly broad. This practically dictates the use of a transmission line of the same characteristic impedance as that shown at the antenna input terminals, or approximately 700 to 800 ohms. Data for the construction of such lines will be found in Chapter Three. It will be found, however, that the spacing required is rather awkward, and also that rather small wire must be used. Both these considerations are disadvantageous mechanically, and the radiation from the line also tends to be high at 28 Mc. because of the wide spacing.

While the usual matching stub can be used to provide an impedance transformation to more satisfactory line impedances, this limits the operation of the antenna to a comparatively narrow range of frequencies centering about that for which the stub is adjusted. On the whole, the best plan is to connect a 600-ohm line directly to the antenna and accept the small mismatch which results. The operation of the antenna will not be adversely affected, and since the standing-wave ratio is quite low (1.33 to 1) the additional loss over the perfectly-matched condition will be inappreciable even for rather long lines. The chief disadvantage is that at some frequencies a slight readjustment of the coupling to the transmitter may be necessary to maintain constant input.

Multiband Antennas

For operation in a number of bands such as those between 3.5 and 30 Mc. it would be impracticable, for most amateurs, to put up a separate antenna for each band. Nor is it necessary; a dipole, cut for the lowest-frequency band to be used, readily can be operated on higher frequencies if one is willing to accept the fact that such harmonic-type operation leads to a change in the directional pattern of the antenna (see Chapter Two), and if one is willing to use "tuned" feeders. A center-fed single-wire antenna can be made to accept power and radiate it with high efficiency on any frequency higher than its fundamental resonant frequency and, with some reduction in efficiency and band width, on frequencies as low as one-half the fundamental.

In addition, methods have been devised for making a single antenna structure operate on a number of bands while still offering a good match to a transmission line, usually of the coaxial type. It should be understood, however, that a "multiband antenna" is not *necessarily* one that will match a given line on all bands on which it is intended to be used. Even a relatively short whip-type antenna can be operated as a multiband antenna with suitable loading, which may be in the form of a coil at its base on those frequencies where loading is needed, or which may be incorporated in the tuned feeders which run from the transmitter to the base of the antenna.

This chapter describes a number of systems that can be used on two or more bands. Beam antennas are not included, these being treated separately in later chapters.

DIRECTLY FED ANTENNAS

The simplest multiband antenna is a random length of No. 12 or No. 14 wire. Power can be fed to the wire on practically any frequency by one or the other of the methods shown in Fig. 6-1. If the wire is made either 67 or 135 feet long, it can also be fed through a tuned circuit as in Fig. 6-2. It is advantageous to use an s.w.r. bridge or other indicator in the coax line at the point marked "X".

If a 28- or 50-Mc. rotary beam has been installed, in many cases it will be possible to use the beam's feed line as an antenna on the lower frequencies. Connecting the two wires of the feeder together at the station end will give a random-length wire that can be conveniently coupled to the transmitter as in Fig. 6-1. The rotary system at the far end will serve only to

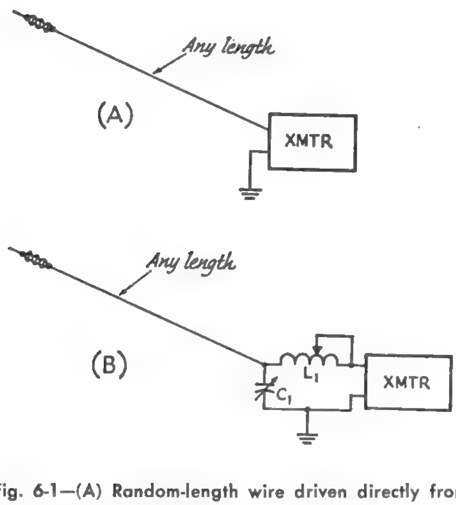


Fig. 6-1—(A) Random-length wire driven directly from the pi-network output of a transmitter. (B) L network for use in cases where sufficient loading cannot be obtained with (A). C_1 should have about the same plate spacing as the final tank capacitor; a maximum capacitance of 100 pf. is sufficient if L_1 is 20 to 25 μh. A suitable coil would consist of 30 turns of No. 12, 2½ inches in diameter, 6 turns per inch. Bare wire should be used so the tap can be placed as required for loading the transmitter.

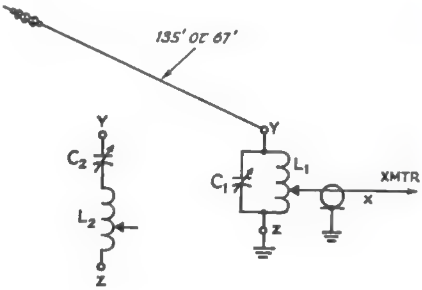


Fig. 6-2—If the antenna length is 135 feet, a parallel-tuned coupling circuit can be used on each amateur band from 3.5 through 30 Mc. C_1 should duplicate the final-tank tuning capacitor and L_1 should have the same dimensions as the final-tank inductor on the band being used. If the wire is 67 feet long series tuning can be used on 3.5 Mc. as shown at the left; parallel tuning will be required on 7 Mc. and higher-frequency bands. C_2 and L_2 will in general duplicate the final tank tuning capacitor and inductor, the same as with parallel tuning.

The *L* network shown in Fig. 6-1B is also suitable for these antenna lengths.

"end-load" the wire and will not have much other effect.

One disadvantage of all such directly fed systems is that part of the antenna is practically within the station, and there is a good chance that trouble with r.f. feedback will be encountered. The r.f. within the station can often be minimized by choosing a length of wire so that a *current loop* occurs at or near the transmitter. This means using a wire length of a quarter wavelength (65 feet at 80 meters, 33 feet at 40 meters), or an odd multiple of a quarter wavelength (¾ wavelength is 195 feet at 80 meters, 100 feet at 40 meters). Obviously this can be done for only one band, in the case of even harmonically related bands, since the wire length that gives a current loop at the transmitter will give a voltage loop at two (or four) times that frequency.

When operating with a random-wire antenna, as in Figs. 6-1 and 6-2, it is wise to try different types of grounds on the various bands, to see which will give the best results. In many cases it will be satisfactory to return to the transmitter chassis for the ground, or directly to a convenient water pipe. If neither of these works well (or the water pipe is not available), a length of No. 12 or No. 14 wire (approximately ¼ wavelength long) can often be used to good advantage. Connect the wire at the point in the circuit that is shown grounded, and run it out and down the side of the house, or support it a few feet above the ground if the station is on the first floor or in the basement. It should not be connected to actual ground at any point.

END-FED ANTENNAS

When a straight-wire antenna is fed at one end by a two-wire line, the length of the antenna portion becomes fairly critical if radiation from the line is to be held to a minimum. Such an antenna system for multiband operation is the "end-fed" or "Zepp-fed" antenna shown in Fig. 6-3. The antenna length is made a half wavelength at the lowest operating frequency. The feeder length can be anything that is convenient, but feeder lengths that are multiples of a quarter wavelength generally give trouble with parallel currents and radiation from the feeder portion of the system. The feeder length given in Table 6-I is free of this trouble. The feeder can be an open-wire line of No. 14 solid copper wire spaced 4 or 6 inches with ceramic or plastic spacers. Open-wire TV line (not the type with a solid web of dielectric) is a convenient type to use. This type of line is available in approximately 300- and 450-ohm characteristic impedance.

The series- and parallel-connected coupling circuits shown in Fig. 6-3 can be replaced by a tapped matching circuit such as is shown in Fig. 6-7.

If one has room for only a 67-foot flat top and yet wants to operate in the 3.5-Mc. band, the two feeder wires can be tied together at the transmitter end and the entire system treated as a random-length wire fed directly, as in Fig. 6-1.

While it is generally sufficient to cut the antenna to length by tape measurement, a closer check can be obtained by inserting an r.f. ammeter in the feeders at A and B in Fig. 6-3. If two similar meters are not available, one meter can be moved from one feeder to the other, closing the circuit in the unmetered feeder with a short jumper wire. When the antenna length is correct, the meter readings in both feeders

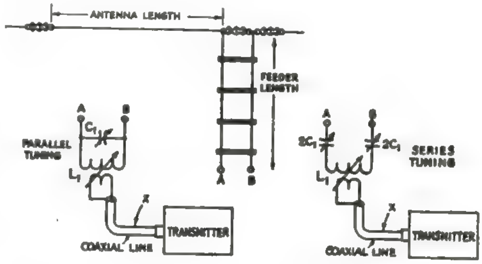


Fig. 6-3—The end-fed antenna using tuned feeders will require a series- or parallel-tuned coupling circuit, depending upon the feeder length and the band (see Table 6-I). L_1C_1 should tune to the operating frequency. For the parallel-tuned circuit, the capacity C_1 should be at least 100 pf. for 3.5 Mc., 50 for 7 Mc. and 25 for 14 Mc. In the series-tuned circuit, the values of 2 C_1 will be considerably lower (roughly one-half to one-fourth), and consequently L_1 will be larger. A single series-tuning capacitor can be used, of a value equal to half of 2 C_1.

Point "X" shows where an s.w.r. indicator can be inserted; a current indicator or meter can be used at A or B.

will be the same, provided there are no parallel currents on the feedline. The simplest insurance against parallel currents is to use a feeder length that is not a multiple of a quarter wavelength, e.g. the 45-foot length recommended in Table 6-I. The flat top can be made long at first and trimmed at the far end until a balanced condition is obtained at the favorite operating

TABLE 6–I

End-Fed Antennas Using Tuned Lines

The antenna lengths are approximate and can be trimmed to the favorite operating frequency by the method given in the text.

Antenna Length (feet)	*Feeder Length (feet)*	*Band (Mc.)*	*Type of Coupling (Circuit)*
135	45	3.5, 7, 14, 21 28	Series Parallel
67	45	7, 14, 21 28	Series Parallel

frequency. It will be found that the currents will balance over only a narrow range of frequencies, because of the changes in current distribution with frequency.

CENTER-FED ANTENNAS

The simplest and most flexible all-band antennas are those using open-wire parallel-conductor feeders to the center of the antenna, as in Fig. 6-4. Because each half of the flat top is the same length, the feeder currents will be balanced at all frequencies unless, of course, unbalance is introduced by one half of the antenna being closer to ground (or a grounded object) than the other. For best results and to maintain feed-current balance, the feeder should run away at right angles to the antenna, preferably for at least a quarter wavelength.

Center feed is not only more desirable than end feed because of inherently better balance, but generally also results in a lower standing-wave ratio on the transmission line provided a parallel-conductor line having a characteristic impedance of 450 to 600 ohms is used. TV-type open-wire line is satisfactory for all but possibly high-power installations (over 500 watts), where heavier wire and wider spacing is desirable to handle the larger currents and voltages.

The length of the antenna is not critical nor is the length of the line. However, some combinations will be less prone than others to have parallel line currents (see Chapter 3). From this standpoint the lengths shown in Table 6-II are recommended, but they are not the only effective ones.

The coupling circuits shown in Fig. 6-4 and Table 6-II correspond to those shown in Fig. 6-3 and Table 6-I. They are more or less traditional with this type of antenna-feeder system, but have no advantages over the matching circuit discussed at some length in Chapter 3 and shown in constructional form in Figs. 6-6 and 6-7. The latter circuit is recommended for all-around use.

The coupling circuits in Table 6-II will, in general, more or less duplicate the *LC* combination used in the final amplifier tank circuit, for the parallel-tuned cases. The series-tuned coupling circuits may require a larger inductance and a smaller capacitance.

When the antenna length is made less than a half wavelength at the lowest operating frequency, there will be a slight reduction in effectiveness at this frequency, but it is not too

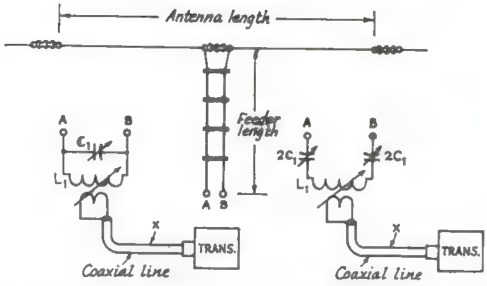


Fig. 6-4—The center-fed all-band antenna requires parallel or series tuning of the coupling circuit, depending upon the system dimensions and the frequency band (see Table 6-II). L_1C_1 should resonate in the range of the operating frequency. For greatest ease of coupling, a low ratio of L_1 to C_1 should be used with parallel tuning, and a high ratio with series tuning.

An s.w.r. bridge can be used at point "X" and a current indicator at A or B.

TABLE 6-II
Center-Fed Antennas Using Tuned Lines

Antenna Length (feet)	*Feeder Length (feet)*	*Band (Mc.)*	*Type of Coupling Circuit*
135	42	3.5–21 28	Parallel Series
135	77½	3.5–28	Parallel
67	42½	3.5 7–28	Series Parallel
67	65½	3.5, 14, 28 7, 21	Parallel Series

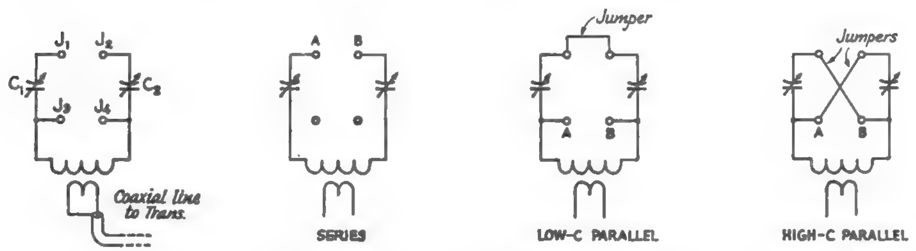


Fig. 6-5—Wiring diagram of a series-parallel type antenna coupler for tuned lines. For 3.5- to 30-Mc. operation, C_1 and C_2 are 75-pf. capacitors with plate spacing equal to that used in the final-amplifier tank tuning capacitor. J_1 through J_4 are banana-plug jacks mounted on the capacitors or on standoff insulators. The coil L_1 is a transmitting type similar in inductance and power rating to that used in the output amplifier. By using jumpers between the jacks, the three types of circuits shown can be obtained. The feeders connect to A-B. To maintain balance to ground, the settings of C_1 and C_2 should be held the same. Insulated extension shafts on the capacitors should be used to eliminate hand-capacitance effects while tuning.

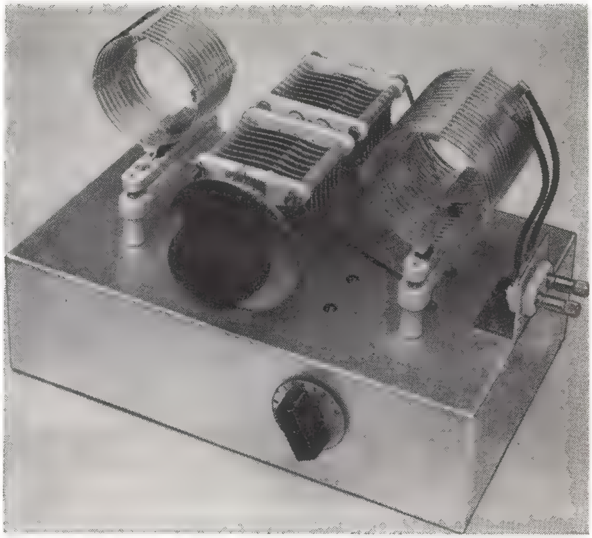


Fig. 6-6—Matching circuit for coupling balanced line to a coaxial link. It may also be used between two coaxial lines as described in the text. The coil at the left is simply "stored" on the chassis as a convenience for changing between two favorite bands. A "Monimatch" bridge is mounted under the 7 × 11 × 3-inch chassis. This coupler will easily handle power levels up to 500 watts.

important until the length is made a quarter wavelength or less.

FEEDER-TO-TRANSMITTER MATCHING CIRCUITS FOR MULTIBAND ANTENNAS

When commercially available link-type coils are used in the series- and parallel-tuned coupling circuits shown with the antenna drawings in Figs. 6-3 and 6-4, it is often difficult to get sufficient coupling to load the transmitter properly. Also, in the case where an s.w.r. bridge is used for adjustment, it may be impossible to find a set of adjustments that will give a low standing-wave ratio in a coaxial line between the transmitter and the coupler.

With transmitters designed to work into resistive loads of 50 to 75 ohms, which is the case with most transmitters of current design, it is highly desirable to use a coupler or matching circuit capable of matching a coaxial line to a wide range of line input impedances. The matching circuit shown in Figs. 6-6 to 6-9, inclusive, is such a unit. Connections for both balanced and unbalanced lines are shown in Fig. 6-7; unbalanced lines are used with some of the antenna systems described later in this chapter. The operating principles and method of adjustment of the system are given in Chapter Three.

The coupling between L_1 and L_2 is adjusted

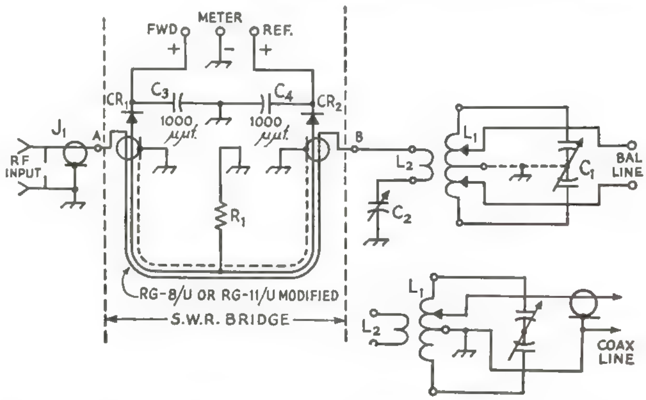


Fig. 6-7—Circuit of the coax-coupled matching circuit. The s.w.r. bridge, a highly useful aid in adjustment, may be omitted if desired, in which case points *A* and *B* are simply connected together.

C_1—100 pf. per section variable, 0.075-inch spacing (Johnson 154-505).
C_2—700 to 800 pf.; dual-section 365- to 400-pf. broadcast-receiver type capacitor with sections in parallel.
C_3, C_4—0.001-μf. disk ceramic.
CR_1 CR_2—1N34A or equivalent.
J_1—Coax receptacle, chassis-mounting type.
L_1, L_2—See table at right.
R_1—See text.

Coil Data

Band, Mc.	L_1 Turns	Wire Size	Dia., In.	Turns/In.	L_2 Turns	Wire Size	Dia., In.	Turns/In.
3.5	44	16	2½	10	10	16	2	10
7	18	12	2½	6	6	16	2	10
14	10	12	2½	6	3	16	2	10
21-28	6	12	2½	6	2	16	2	10

L_1 for 3.5 Mc. is made from Air-Dux 2010; all other bands, B. & W. 3905-1 or Air Dux 2006. L_2 is B. & W. 3907-1 or Air Dux 1610.

Fig. 6-8—Below-chassis view of the matching circuit, showing Monimatch made from a section of coax cable. The crystal rectifiers are mounted on dual tie-point strips, with R_1 between them.

by means of a variable capacitor, C_2, to facilitate matching between the main transmission line from the antenna and the coax line to the transmitter. The coils are constructed from commercially available coil material, and the link (L_2) inductances are chosen to provide adequate coupling for flat lines. The link coil, of smaller diameter than the tank coil L_1, is mounted inside the latter at the center. Duco cement is used to hold the coils together at their bottom tie strips. The coils are mounted on Millen type 40305 plugs and require no other support than the stiffness of the short lengths of wire going into the end prongs of the plug from the tank coil. Short lengths of spaghetti tubing are slipped over the leads to the link coil where they go between the tank coil turns to reach the plug.

Taps on the tank coil for connection to a paralled-conductor transmission line are made by means of clips. If coils are changed frequently it will be convenient, after finding the proper tap points for each band, to bend ordinary soldering lugs around the wire and solder them in place so they project radially from the coil. Used this way, the clips provide an easy and rapid method of connecting and disconnecting the line.

Monimatch

The circuit includes a directional coupler of the Monimatch type to assist in adjusting the circuit to match the coax line. It is constructed from a 24-inch length of either RG-8/U or RG-11/U (depending on the Z_0 of the coax line between the transmitter and the matching circuit). The pickup line, to which R_1 and the crystal rectifiers are connected, is a length of No. 30 enameled wire inserted between the insulation and the shield-braid outer conductor of the coax cable, using the method described in Chapter Three. In constructing this line section be careful not to scrape the enamel from the wire, and after the braid has been smoothed out to its original length check between it and the pickup wire with an ohmmeter to make sure the two are not short-circuited.

The cable is formed into a double turn, as shown in Fig. 6-8, so that the center, where R_1 connects to the pickup wire, is close to the ends. This keeps the ground paths to minimum length and helps in obtaining proper balance in the bridge. The braided outsides of the turns are spot soldered together at several points to reduce the effect of unwanted currents on the surface, and also to improve the assembly mechanically.

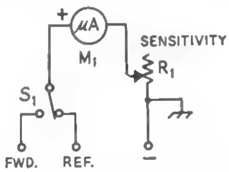


Fig. 6-9—Indicator-unit circuit. For low power and low frequencies, M_1 should be a 0-100 microammeter. A 0-1 milliammeter will suffice in other cases.

R_1—25,000-ohm composition control.

S_1—S.p.d.t. toggle.

Fig. 6-9 shows a suitable circuit for the null indicator for the Monimatch. This can be assembled in any fashion that is convenient, and can be installed wherever desired since the leads between it and the Monimatch carry only low-current d.c.

Adjusting the Monimatch is simply a matter of finding the value of R_1 that gives a good null reading with the indicating meter connected to the "reflected" position when the output end is terminated in a resistive load of either 52 or 75 ohms, depending on whether RG-8/U or RG-11/U is used. If a suitable dummy load is available the wiring to L_2 should be disconnected at B in Fig. 6-7 and the dummy load connected between B and ground (that is, to the output terminals of the Monimatch). R_1 may be set to the proper value by trying several values of half-watt carbon resistors, or combinations in parallel, to find the resistance that gives the deepest null. A value of about 35 ohms proved to be optimum with RG-8/U in the bridge shown in the photograph.

Alternatively, a dummy load may be con-

nected to the balanced line terminals, and the Monimatch disconnected at *B*. If a suitable bridge can be borrowed, it can be connected at *B* and r.f. power fed through it to the matching circuit, which should then be adjusted to match the coax line. This establishes a load of known value which may then be used for adjustment of the built-in Monimatch as described above, after the connection at *B* has been restored.

Matching-Circuit Adjustment

The method of adjusting a matching circuit of this type has been described in Chapter Three. The construction is such that either the center tap of L_1 or the rotor of C_1 may be grounded to the chassis, since C_1 is mounted on small standoff insulators. Insofar as normal balanced-line operation is concerned, it makes no difference which is grounded (or neither). Grounding will, however, affect any parallel or "antenna" currents on the line. In general, the effect of such currents will be minimized if the ground connection showing the least r.f. current is chosen. This test should also be tried with and without an actual earth connection to the matching-circuit chassis.

The coupler may be used between coaxial lines by grounding the center tap of L_1 and connecting the outer braid of the coax line to the chassis and the inner conductor to a single tap on the coil as shown in Fig. 6-7. The method of adjustment is otherwise the same as for balanced lines.

In general, the tuning will be less critical, and the circuit will work over a wider frequency range without readjustment, if the taps are kept as far toward the ends of the coil as possible and C_2 is set at the largest capacitance that will permit bringing the s.w.r. in the coax link down to 1 to 1.

OFF-CENTER FEED—THE "WINDOM" ANTENNA

A multiband antenna that enjoyed considerable popularity in the 1930s is the "off-center feed" or "Windom," named after the amateur who wrote a comprehensive article about it. Shown in Fig. 6-10A, it consists of a half-wavelength antenna on the lowest-frequency band to be used, with a *single-wire* feeder connected off center as shown. The antenna will operate satisfactorily on the even-harmonic frequencies, and thus a single antenna can be made to serve on the 80-, 40-, 20- and 10-meter bands. The single-wire feeder shows an impedance of approximately 600 ohms to ground, and since the return circuit for the feed system is through the earth, a good ground connection is important to the effective operation of the antenna. Also, the system works best when installed over ground having high conductivity.

Theoretically the single-wire feeder can be any convenient length, since its characteristic impedance is matched by the antenna impedance at the point where the feeder is connected. However, this type of feeder is susceptible to

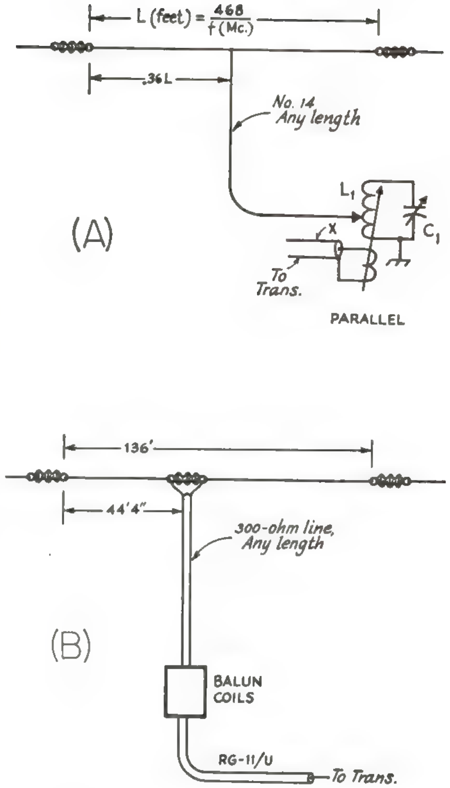


Fig. 6-10—Two versions of the off-center-fed antenna. (A) Single-wire feed. The single-wire feeder can be connected directly to the "hot" output terminal of a pi network in the transmitter. Alternatively, the link-coupled circuit shown may be used, with a separate ground connection as indicated; this type of coupling helps reduce troubles from r.f. currents on the station equipment. The matching circuit described earlier in this chapter also can be used, with unbalanced output. Both circuits should be adjusted as described for the tapped matching circuit. (B) Two-wire feed, using 300-ohm TV line. The balun coils may be omitted and the 300-ohm line connected directly to the output terminals of a pi network in the transmitter, but this is not recommended because it leads to r.f. troubles of the type described in the text. The matching circuit shown earlier in this chapter may be substituted for the balun coils if desired.

parallel-type currents just as much as the two-conductor type (see Chapter Three), and some feeder lengths will lead to "r.f. in the shack" troubles, especially when the feeder goes directly to a pi network in the transmitter. Adding or subtracting ⅛ wavelength or so of line usually will help cool things off in such cases.

A newer version of the off-center-fed antenna (miscalled "Windom") uses 300-ohm TV Twin-Lead instead of a single-wire line. This system is shown in Fig. 6-10B. The claim has been

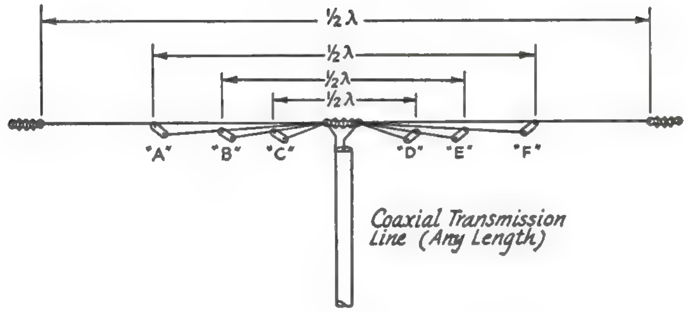


Fig. 6-11—Multiband antenna using parallelied dipoles all connected to a common low-impedance transmission line. The half-wave dimensions may be either for the centers of the various bands or selected to fit favorite frequencies in each band. Length of half wave in feet is 468/frequency in Mc.

made that the 300-ohm line is matched by the antenna impedance at the connection point both on the antenna's fundamental frequency and on harmonics, but there is little theoretical justification for this. The system is particularly susceptible to parallel line currents because of the unsymmetrical feeder connection, and probably in many cases the line acts more like a single-wire feeder than a parallel-conductor one. The parallel currents on the line can be choked off by using balun coils (see Chapter Three) as shown in Fig. 10B. The same balun will transform the impedance to 75 ohms, in cases where the line actually shows a resistive input impedance of 300 ohms.

With either of the off-center-fed systems the feeder should be brought away from the antenna at right-angles for at least a quarter wavelength before any bends are made. Any necessary bends should be made gradually.

MULTIPLE-DIPOLE ANTENNAS

The antenna system shown in Fig. 6-11 consists of a group of center-fed dipoles all connected in parallel at the point where the transmission line joins them. One such dipole is used for each band on which it is desired to work, and as many as four have been used, as indicated in the sketch. It is not generally necessary to provide a separate dipole for the 21-Mc. band since a 7-Mc. dipole works satisfactorily as a third-harmonic antenna on this band.

Although there is some interaction between the dipoles it has been found in practice that the ones that are not resonant at the frequency actually applied to the antenna have only a small effect on the feed-point impedance of the "active" dipole. This impedance is therefore approximately that of a single dipole, or in the neighborhood of 60-70 ohms, and the system can be fed through a 50- or 75-ohm line with a satisfactorily low standing-wave ratio on the line.

Since the antenna system is balanced, it is desirable to use a balanced transmission line to feed it. The most desirable type of line is 75-ohm solid-dielectric Twin-Lead. The transmitting variety of line should be used, since the 75-ohm receiving-type line has rather high loss, even when matched. However, either 52-ohm or 75-ohm coaxial line can be used; coax line introduces some unbalance, but this is not intolerable on the lower frequencies.

The separation between the dipoles for the various frequencies does not seem to be especially critical, so far as experience indicates. One set of wires can be suspended from the next larger set, using insulating spreaders (of the type used for feeder spreaders) to give a separation of a few inches.

An interesting method of construction used successfully by ON4UF is shown in Fig. 6-12. The antenna has four dipoles (for 7, 14, 21 and 28 Mc.) constructed from 300-ohm ribbon transmission line. A single length of ribbon makes two dipoles. Thus, two lengths, as shown in the sketch; serve to make dipoles for four bands. Ribbon with copper-clad steel conductors (Amphenol type 14-022) should be used because all of the weight, including that of the feed line, must be supported by the uppermost wire.

Two pieces of ribbon are first cut to a length suitable for the two halves of the longest dipole. Then one of the conductors in each piece is cut to proper length for the next band higher in

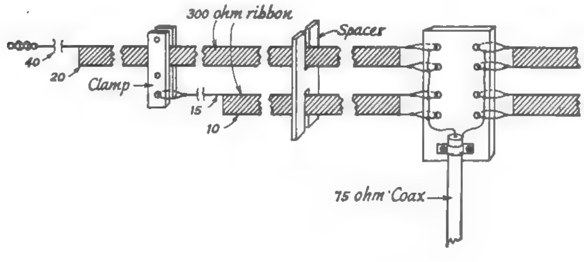


Fig. 6-12—Sketch showing how the Twin-Lead multiple-dipole antenna system is assembled. The excess wire and insulation are stripped away.

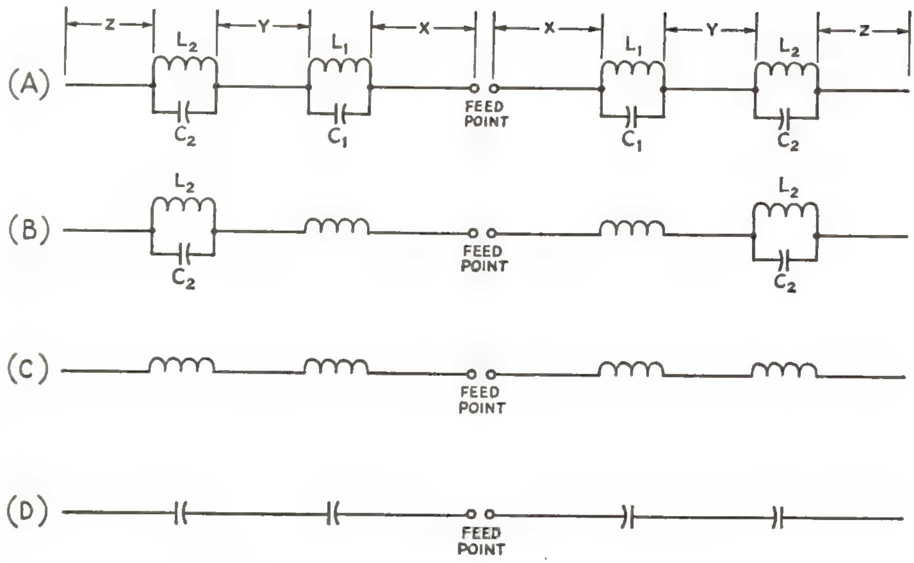


Fig. 6-13—Development of the "trap" dipole for operation on fundamental-type resonance in several bands.

frequency. The excess wire and insulation is stripped away. A second pair of lengths is prepared in the same manner, except that the lengths are appropriate for the next two higher frequency bands.

A piece of thick polystyrene sheet drilled with holes for anchoring each wire serves as the central insulator. The shorter pair of dipoles is suspended the width of the ribbon below the longer pair by clamps also made of poly sheet. Intermediate spacers are made by sawing slots in pieces of poly sheet so that they will fit the ribbon snugly.

Dimensions, as determined by use of a grid-dip oscillator, are shown in Table 6-III. The system showed an impedance of close to 70 ohms on all bands, and the s.w.r. on a 75-ohm line was low and nearly constant.

Table 6–III Twin-Lead Parallel-Dipole Antenna Dimensions

Frequency (Mc.)	*Length Each Half*		
	Meters	*Feet*	*In.*
7.1	9.95	32	8
14.1	4.60	15	1
21.2	3.44	11	2
28.2	2.34	7	8

TRAP DIPOLES

By using tuned circuits of appropriate design strategically placed in a dipole, the antenna can be made to show what is essentially fundamental resonance at a number of different frequencies. The general principle is illustrated by Fig. 6-13. The two inner lengths of wire, *X*, together form a simple dipole resonant at the highest band desired, say 14 Mc. The tuned circuits L_1C_1 are also resonant at this frequency, and when connected as shown offer a very high impedance to r.f. current of that frequency which may be flowing in the section *X-X*. Effectively, therefore, these two tuned circuits act as insulators for the inner dipole, and the outer sections beyond L_1C_1 are inactive.

However, on the next lower frequency band, say 7 Mc., L_1C_1 shows an inductive reactance and is the electrical equivalent of a coil. If the two sections marked *Y* are now added and their length adjusted so that, together with the loading coils represented by the inductive reactance of L_1C_1, the system out to the ends of the *Y* sections is resonant at 7 Mc., this part of the antenna is equivalent to a loaded dipole on 7 Mc. and will exhibit about the same impedance at the feed point as a simple dipole for that band. The tuned circuit L_2C_2 is resonant at 7 Mc. and acts as a high impedance for this frequency, so the 7-Mc. dipole is in turn insulated, for all practical purposes, from the remaining outer parts of the antenna.

Carrying the same reasoning one step farther, L_2C_2 shows inductive reactance on the next lower-frequency band, 3.5 Mc., and is equivalent to a coil on that band. The length of the added sections, Z-Z, is adjusted so that, together with the two sets of equivalent loading coils now indicated in C, the whole system is resonant as a loaded dipole on 3.5 Mc. A single transmission line having a characteristic impedance of the same order as the feed-point impedance of a simple dipole can be connected at the center of the antenna and will be satisfactorily matched on all three bands, and so will operate at a low

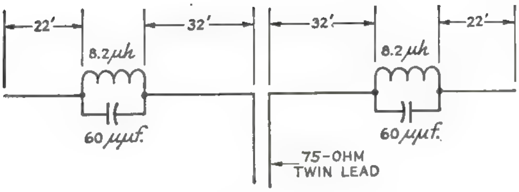


Fig. 6-14—Five-band (3.5 through 28 Mc.) trap dipole for operation with 75-ohm feeder at low s.w.r. (W3-DZZ). The balanced (parallel-conductor) line indicated is desirable, but 75 ohm coax can be substituted with some sacrifice of symmetry in the system. Dimensions given are for resonance (lowest s.w.r.) at 3750, 7200, 14,150, and 29,500 kc. Resonance is very broad on the 21-Mc. band, with s.w.r. less than 2:1 throughout the band.

s.w.r. on all three. A line of 75-ohm impedance is satisfactory; coax may be used, but Twin-Lead will maintain better balance in the system since the antenna itself is symmetrical.

Trap Losses

Since the tuned circuits have some inherent losses the efficiency of this system depends on the Qs of the tuned circuits. Low-loss (high-Q) coils should be used, and the capacitor losses likewise should be kept as low as possible. With tuned circuits that are good in this respect—comparable with the low-loss components used in transmitter tank circuits, for example—the reduction in efficiency as compared with the efficiency of a simple dipole is small, but tuned circuits of low Q can lose an appreciable portion of the power supplied to the antenna.

Dimensions

The lengths of the added antenna sections, Y and Z in the example, must in general be determined experimentally. The length required for resonance in a given band depends on the length/diameter ratio of the antenna conductor and on the L/C ratio of the trap acting as a loading coil. The effective reactance of an LC circuit on half the frequency to which it is resonant is equal to ⅔ the reactance of the inductance at the resonant frequency. For example, if L_1C_1 resonates at 14 Mc. and L_1 has a reactance of 300 ohms at 14 Mc., the inductive reactance of the circuit at 7 Mc. will be equal to ⅔ × 300 = 200 ohms. The added antenna section, Y, would have to be cut to the proper length to resonate at 7 Mc. with this amount of loading. Since any reasonable L/C ratio can be used in the trap without affecting its performance materially at its resonant frequency, the L/C ratio can be varied to control the added antenna length required. The added section will be shorter with high-L trap circuits and longer with high-C traps.

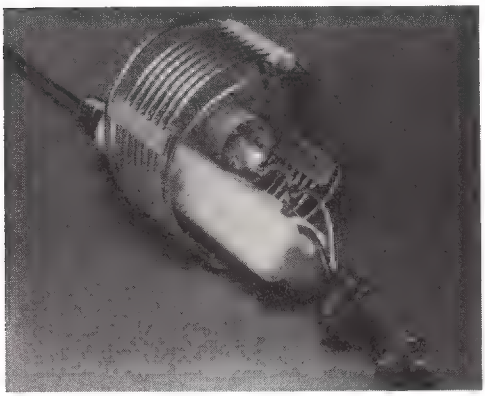


Fig. 6-15—Easily constructed trap for wire antennas (W2CYK). The ceramic insulator is 4¼ inches long (Birnbach 668). The clamps are small service connectors available from electrical-supply and hardware stores (Burndy KS90 Servits).

Higher Frequencies

On bands higher than that for which the inner dipole is resonant all traps in the system show capacitive reactance. Thus at such frequencies the antenna has the equivalent circuit shown at D in Fig. 6-13. The capacitive reactances have the effect of raising the resonant frequency of the system as compared with a simple dipole of the same over-all length.

This effect is greatest near the resonant frequency of the inner dipole X-X and becomes less marked as the frequency is increased, since the capacitive reactance decreases with increasing frequency. The system therefore can be used on higher-frequency bands as a harmonic-type antenna, but obtaining resonance with low impedance will require careful balancing of the trap L/C ratios and the lengths of the various antenna sections.

Five-Band Antenna

One such system has been worked out by W3DZZ for the five amateur bands from 3.5 to 30 Mc. Dimensions are given in Fig. 6-14. Only one set of traps is used, resonant at 7 Mc. to isolate the inner (7-Mc.) dipole from the outer sections, which cause the over-all system to be resonant in the 3.5-Mc. band. On 14, 21 and 28 Mc. the antenna works on the capacitive-reactance principle just outlined. Using 75-ohm Twin-Lead as a feeder, the s.w.r. with this antenna was under 2 to 1 throughout the three high-frequency bands, and the s.w.r. was comparable with that obtained with similarly-fed simple dipoles on 7 and 3.5 Mc.

Trap Construction

Traps frequently are built with coaxial aluminum tubes (usually with polystyrene tubing between them for insulation) for the capacitor, and with the coil either self-supporting or wound on a form of larger diameter than the tubular capacitor. The coil is then mounted coaxially

with the capacitor to form a unit assembly that can be supported at each end by the antenna wires. Examples of this type of construction are shown in Chapter 12.

A simpler type of trap, easily assembled from readily available components, is shown in Fig. 6-15. A small transmitting-type ceramic capacitor is used, together with a length of commercially available coil material, these being supported by an ordinary antenna strain insulator. The circuit constants and antenna dimensions differ slightly from those of Fig. 6-14, in order to bring the antenna resonance points closer to the centers of the various phone bands. Construction data are given in Fig. 6-16. If a 10-turn length of inductor is used, a half-turn from each end may be used to slip through the anchor holes in the insulator to act as leads.

The components used in these traps are sufficiently weatherproof in themselves so that no additional treatment for this purpose has been found to be necessary. However, if it is desired to protect them from the accumulation of snow or ice a plastic cover can be made by cutting two disks of polystyrene slightly larger in diameter than the coil, drilling at the center to pass the antenna wires, and cementing a plastic cylinder on the edges of the disks. The cylinder can be made by wrapping two turns or so of 0.02-inch poly or Lucite sheet around the disks, if no suitable ready-made tubing is available.

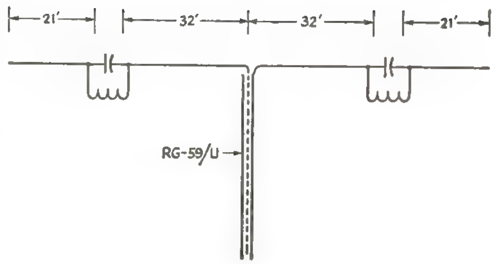


Fig. 6-16—Layout of multiband antenna using traps constructed as shown in Fig. 6-15. The capacitors are 100 pf. each, transmitting type, 5000-volt d.c. rating (Centralab 850SL-100N). Coils are 9 turns No. 12, 2½ inch diameter, 6 turns per inch (B & W 3905-1 or Air Dux 2006), with end turns spread as necessary to resonate the traps to 7200 kc. These traps, with the wire dimensions shown, resonate the antenna at approximately the following frequencies on each band: 3900, 7250, 14,100, 21,500 and 29,900 kc. (based on measurements by W9YJH).

Four-Band Trap Dipole

In case there is not enough room available for erecting the 100-odd-foot length required for the five-band antennas just described, Fig. 6-17 shows a four-band dipole operating on the same principle that requires only half the linear space. The trap construction is the same as shown in Fig. 6-15. With the dimensions given in Fig. 6-17 the resonance points are 7200, 14,100, 21,150 and 28,400 kc. The capacitors are 25-pf. transmitting-type ceramic (Centralab type 857). The inductors are 9 turns of No. 12, 2½ inches in diameter, 6 turns per inch (B & W 3905-1), adjusted so that the trap resonates at 14,100 kc. before installation in the antenna.

VERTICAL ANTENNAS

A short vertical antenna can be operated on several bands by loading it at the base, the general arrangement being similar to Figs. 6-1 and 6-2. That is, for multiband work the vertical can be handled by the same methods that are used for random-length wires.

However, a vertical antenna should not be longer than about ¾ wavelength at the highest frequency to be used, if low-angle radiation is wanted. If the antenna is to be used on 28 Mc. and lower frequencies, therefore, it should not be more than approximately 25 feet high, and the shortest possible ground lead should be used. If the base of the antenna is well above actual ground, the ground lead should run to the nearest water or heating pipe.

Another method of feeding is shown in Fig. 6-18. L_1 is a loading coil of adjustable inductance so the antenna can be tuned to resonate on the desired band. It is tapped for adjustment of tuning, and a second tap permits using the coil as a transformer for matching a coax line to the transmitter. Capacitor C_1 is not strictly necessary, but may be helpful on the lower frequencies, 3.5 and 7 Mc., if the antenna is quite short. In that case C_1 makes it possible to tune to resonance with a coil of reasonable dimensions at L_1. C_1 may also be useful on other bands as well, if the system cannot be matched to the coax line with the coil alone.

The coil and capacitor preferably should be installed at the base of the antenna, but if this cannot be done a wire can be run from the antenna base to the nearest convenient location for mounting L_1 and C_1. The extra wire will of course be a part of the antenna, and since it

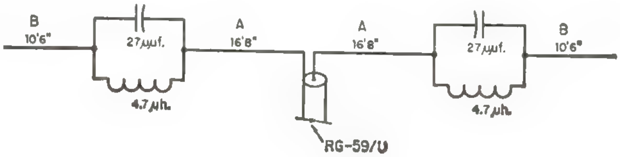


Fig. 6-17—Sketch showing dimensions of a trap dipole covering the 40- through 10-meter bands (K2GU).

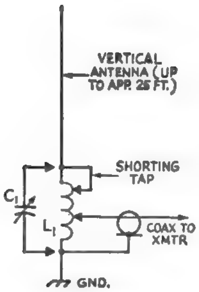


Fig. 6-18—Multiband vertical antenna system using base loading for resonating on 10 to 80 meters. L_1 should be wound with bare wire so it can be tapped at every turn, using No. 12 wire. A convenient size is 2½ inches in diameter, 6 turns per inch (such as B. & W. 3905-1 or Air Dux 2006). Number of turns required depends on antenna and ground lead length, more turns being required as the antenna and ground lead are made shorter. For a 25-foot antenna and a ground lead of the order of 5 feet, L_1 should have about 30 turns. The use of C_1 is explained in the text. The smallest capacitance that will permit matching the coax cable should be used; a maximum capacitance of 100 to 150 pf. will be sufficient in any case.

may have to run through unfavorable surroundings it is best to avoid its use if at all possible.

This system is best adjusted with the help of an s.w.r. indicator. Connect the coax line across a few turns of L_1 and take a trial position of the shorting tap. Measure the s.w.r., then try various positions of the shorting tap until the s.w.r. reaches its lowest value. Then vary the line tap similarly; this should bring the s.w.r. down to a low value. Small adjustments of both taps then should reduce the s.w.r. to close to 1 to 1. If not, try adding C_1 and go through the same procedure, varying C_1 each time a tap position is changed.

Trap Verticals

The trap principle described in Fig. 6-13 for center-fed dipoles also can be used for vertical antennas. There are two principal differences: Only one half of the dipole is used, the ground connection taking the place of the missing half, and the feed-point impedance is one half the feed-point impedance of a dipole. Thus it is in the vicinity of 30 ohms (plus the ground-connection resistance), so 52-ohm cable should be used since it is the type that comes closest to matching.

As in the case of any vertical antenna, a good ground is essential, and the ground lead should be short. Some amateurs have reported successfully using a ground plane dimensioned for the lowest frequency to be used; for example, if the lowest frequency is 7 Mc. the ground-plane radials can be approximately 34 feet long.

Traps of the type shown in Fig. 6-15 can be used in a vertical wire suspended from a support, but the type of trap shown in Chapter 12 for rotatable beam antennas should be used in tubing verticals.

HARMONIC RADIATION FROM MULTIBAND ANTENNAS

Since a multiband antenna is intentionally designed for operation on a number of different frequencies, any harmonics or spurious frequencies that happen to coincide with one of the antenna's resonant frequencies will be radiated with very little, if any, attenuation. Particular care should be exercised, therefore, to prevent such harmonics from reaching the antenna.

Multiband antennas using tuned feeders have a certain inherent amount of built-in protection against such radiation since it is nearly always necessary to use a tuned coupling circuit between the transmitter and the feeder. This adds considerable selectivity to the system and helps to discriminate against all frequencies other than the desired one.

Multiple dipoles and trap antennas do not have this feature since the object in design is to make the antenna show as nearly as possible the same resistive impedance in all the amateur bands the antenna is supposed to cover. It is advisable to conduct tests with other amateur stations to determine whether harmonics of the transmitting frequency can be heard at a distance of, say, a mile or so. If they can, more selectivity should be added to the system since a harmonic that is heard locally, even if weak, may be quite strong at a distance because of propagation vagaries. A matching circuit of the type described earlier in this chapter will add enough selectivity to take care of harmonics of the strength generated by transmitters of good design and construction.

Antennas for 160 Meters

With respect to sky-wave transmission, the requirements that the antenna system must meet on 160 meters do not differ materially from those which hold on the high-frequency bands. Of course, waves entering the ionosphere even vertically are reflected back to earth so that there is no such phenomenon as skip distance on these frequencies. However, it is still true that to cover the greatest possible distance the waves must enter the ionosphere at low angles. Although a given distance may be covered by multiple hops when the radiation angle is high, there will be less absorption, and hence the signal strength will be greater, at the same point when the wave reaches it by only one hop.

On the "160-meter" band the ground wave assumes considerable importance for transmission over short distances. The useful range of the ground wave will depend upon the transmitter power, the background noise at the receiver, and the type of soil over which the wave must travel. If the antenna system radiates most of the transmitter power at relatively low angles, particularly along the ground, the ground wave will give reliable communication over distances from 50 to considerably over 100 miles, the latter distances applying where conditions are particularly favorable, as when the path is mostly over sea water.

POLARIZATION

It was mentioned in Chapter One that a ground wave must be vertically polarized, so that the radiation from an antenna which is to produce a good ground wave likewise must be vertically polarized. This dictates the use of an antenna system the radiating part of which is mostly vertical. A horizontally polarized antenna will produce practically no ground wave, and it is to be expected that such an antenna will be ineffective for daytime communication. This is because absorption in the ionosphere in the daytime is so high at these frequencies that the reflected wave is too weak to be useful. At night a horizontal antenna will give better results since nighttime ionosphere conditions permit the reflected wave to return to earth without excessive attenuation. The difference between daytime and nighttime conditions is similar to that existing on the broadcast band, where distant stations can be heard well at night but not at all in the day.

There is still another reason why a vertical antenna is better than the horizontal for 160-meter work. Comparison of the ground-reflection factors in Figs. 2-26 and 2-27, Chapter Two, for horizontal antennas at heights of ¼ and ⅛ wave will show that at the lower height the ground is less effective in reinforcing radiation. At 160 meters even a height of ⅛ wave, about 65 feet, is not easy for all amateurs to attain, while a height of ¼ wave is out of the question for nearly everyone. Any reasonable height is small in terms of wavelength, so that a horizontal antenna on 160 meters is a poor radiator at angles useful for long distances ("long," that is, for this band). Its chief field of usefulness is for communication over relatively short distances at night.

The chief disadvantage of vertical polarization is the fact that the stronger ground wave is more likely to cause interference with nearby broadcast receivers.

Vertical-Antenna Design Considerations

For good night coverage at distances toward the limit of the ground wave it is desirable to use an antenna that will give comparatively little radiation at angles above about 45 degrees. This is because the high-angle radiation returns to earth within the useful range of the ground wave, and in the outer part of this range may have intensity comparable with that of the ground wave itself. The sky waves arrive at the ground in random phase with respect to the ground wave, giving rise to severe fading in this area. The antenna should, therefore, confine its radiation to angles sufficiently low so that the nearest point to the transmitter at which sky waves return to earth is just beyond the limits of the ground wave.

The various conditions can be met by the use of an antenna a half wave high, but this is impractical since a height of over 250 feet would be required. Fortunately it is possible to approach the effect of a half-wave antenna by suitable treatment of a much lower structure.

A vertical antenna will be most effective when it can be erected in a fairly clear spot so that the ground wave is not absorbed in nearby buildings. Frame buildings are not likely to cause much trouble, but it is best to keep clear of steel structures by at least a wavelength or two.

GROUNDED ANTENNAS

It was explained in Chapter Two that a quarter-wavelength grounded antenna is resonant, and that a still smaller one can be made resonant by "loading" it at the base. However, as pointed

out in that chapter, it is far better to do the loading at the top of the antenna, to raise the point of current maximum in the antenna and thus to increase the radiating efficiency of the system. Several methods for top loading an antenna were described.

Bent Antennas

Perhaps the simplest method of meeting the fundamental requirement of keeping the current loop high is to use a bent antenna such as is shown in Fig. 7-1A, with part of the antenna vertical and part horizontal. The horizontal part should be one quarter-wave in length so that the current loop will appear at the top of the vertical portion. The current distribution will be as shown in the drawing, assuming that the vertical portion is ¼ wave high. If smaller heights are used, the horizontal portion should still be ¼ wave in length. Since the most useful radiation is from the vertical part, it is of course, desirable to make the antenna as high as possible.

The length of the horizontal portion can be calculated from the formula

$$\textit{Length of quarter wave (ft.)} = \frac{234}{f\,(\text{Mc.})}.$$

There is no need for excessive accuracy in determining this length, since a discrepancy of 5 or 10 feet will make comparatively little difference in the performance of the antenna.

The lower end of the antenna is grounded through a loading circuit that tunes the system to resonance and also provides a means for coupling power from the transmitter into the antenna. The constants of the loading circuit will depend upon the total length of the antenna system, and therefore depend upon the antenna height. For heights between 40 and 70 feet a circuit of the type shown in Fig. 7-2 will be suitable, provided the leads between the bottom of the antenna and the coupling circuit, and between the loading circuit and the effective ground, are only a few feet in length. These leads are part of the effective length of the antenna, and must be added to the antenna length in determining the actual constants required in the loading circuit.

For maximum effectiveness, the vertical part of the antenna should actually be vertical, and not simply run off at some convenient angle from the operating room to the top of the pole.

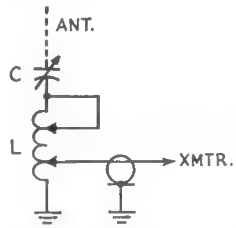


Fig. 7-2—A practical loading and coupling circuit for antennas of the type shown in Fig. 7-1B when the height X is ⅛ wavelength or less (up to 65 feet approximately). The series tuning capacitor C should be 250 to 500 pf.; receiving-type capacitors will suffice for moderate powers. Coil L may consist of 20 turns of No. 12 wire space-wound (6 turns per inch) to a diameter of 3 inches, arranged so that it can be tapped conveniently at least every few turns. Tuning procedure is that for series tuning as described in Chapter Three. An r.f. ammeter may be connected in series with the antenna where it joins C. A 2.5-ampere instrument will suffice for powers up to a few hundred watts.

The wire may come down the pole on standoff insulators, or may be pulled down vertically from the horizontal strain insulator after the fashion shown in Fig. 7-3. Wire guys on the pole should be broken up at intervals of 25 feet or so with egg-type insulators to prevent pickup of r.f. energy from the antenna.

Antennas of this type offer an opportunity for use of a rather simple feeder system that permits installing the antenna at some distance from the transmitter. If the antenna height is ⅛ wave, for example, the total length is ⅜ wave including the horizontal part. An additional ¼ wave wire may be added to the antenna, as shown in Fig. 7-4, to make the total length ½ wave. This extra section is connected to the bottom of the vertical wire and is used as a feeder. It should run parallel to and fairly close to the ground for as much of its length as possible (a height of 7 feet is permissible so that it will not be a hazard to walkers) and terminate at the transmitter in a parallel-tuned circuit, the other end of which is grounded. (The length of the ground lead should be included in the "feeder" length.) At this point the impedance looking into the feeder and antenna has its highest value so that losses in the ground connection are relatively low. There will be very little current in the ground lead under these conditions, but an ammeter inserted at the base of the vertical portion will read about 70 per cent of the current at the top.

Such a "feeder" does comparatively little

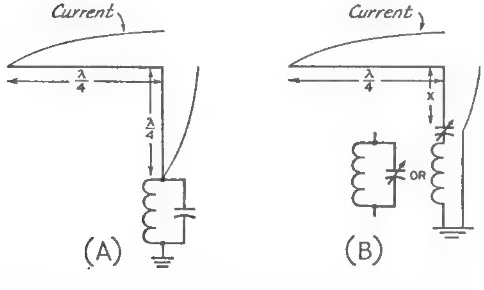


Fig. 7-1—Bent antennas using a quarter-wave horizontal section to bring a current loop at the top of the vertical wire. A quarter-wave vertical section is shown at A; at B the height X is made as great as the circumstances permit. Series tuning may be used for lengths of X up to about ⅛ wavelength; parallel tuning for greater lengths of X.

radiating because it is parallel to and close to the ground and because it represents the section of the antenna which carries the least current. In cases where the antenna height is not an eighth wavelength the "feeder" length, including the ground lead, should be ½ wavelength less the actual length of the antenna from the base to the far end of the horizontal portion. The length of a half wave is given closely enough by the formula

$$\textit{Length of half wave} \text{ (ft.)} = \frac{468}{f\,(\text{Mc.})}.$$

The feeder may be made longer or shorter than the exact length necessary to make the whole system a half wave long, if more convenient, provided the whole system is brought to resonance by means of the coupling system. However, excessive length in a feeder of this type is not desirable. Also, it is preferable to have the length to the ground connection a half wave so that the current in the ground lead will be minimum, which means lowest loss in the ground connection.

Grounds

One of the chief problems of obtaining optimum performance on 160 meters is that of getting a good low-resistance ground. The ideal arrangement is an extensive radial system (see next section) but this is seldom practicable at an amateur station. In locations having a community water-distribution system, probably as good a substitute as any for the radial ground is a connection to a cold-water pipe, since it is part of an extensive network of buried piping. Always select a cold-water pipe since it usually goes more directly to ground than the hot-water variety. Gas pipes never should be used because insulated joints are sometimes included in the piping. Wherever possible, the connection to the cold water piping should be made directly at the point where the pipe enters the ground; that is, on the street side of the water meter. The length of the ground lead necessarily must be taken into account in computing the total length of the antenna.

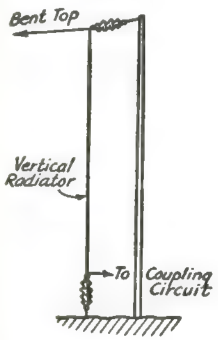


Fig. 7-3—An arrangement for keeping the main radiating portion of the antenna vertical.

To make the connection, carefully clean the pipe by scraping and sandpapering. Fit on a clean ground clamp, make it good and tight, and make sure that the ground wire makes a good electrical connection to the strap. Solder it if necessary. The assembly may be rubber-taped to prevent oxidation if there is considerable dampness.

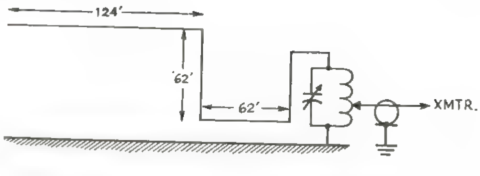


Fig. 7-4—Bent antenna ⅛ wave high, with a "feeder" section making the total length ½ wavelength. The length figures are for 1900 kc., approximately, and the same antenna may be used over the whole band. The parallel-tuned coupling circuit should be capable of being tuned independently to the operating frequency, and the inductance of the coil preferably should be variable by means of taps so that the optimum L/C ratio can be secured.

If it is impossible to reach the pipe at the point where it enters the ground, a connection of the type described above may be made to any convenient cold-water pipe as a secondary resort. In such cases, estimation of the effective length of the ground lead is difficult, since piping systems sometimes are rather extensive and hence have considerable capacitance to ground. The effective length usually will be appreciably less than the actual length of the shortest path which might be traced back to ground along the piping, and in the case of a ground to a heating system may be quite small because of the large masses of metal at the radiators. In such cases the amount of loading for bringing the system to resonance must be determined experimentally.

In locations where there is no such water-supply system a simple outdoor ground may be made by driving a length (6 feet or more) of 1-inch pipe into the soil. If possible, pick a spot where there is considerable natural moisture; the

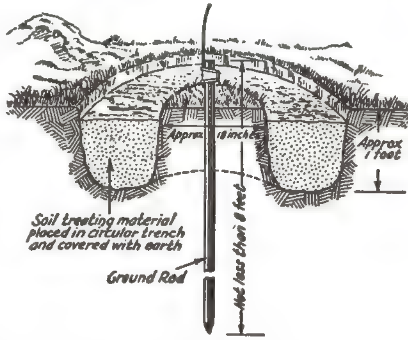


Fig. 7-5—Ground system treated to increase conductivity. The circular trench is filled with rock salt, magnesium sulphate, or copper sulphate, put in dry and then flooded with water. After treatment, the trench is covered with earth. Fifty pounds of treating material so disposed will have a life of two or three years.

resistance will be less under such conditions. Four pipes arranged at the corners of a 10-foot square, all connected together at the top, will be considerably better than one.

A quite good low-resistance ground connection can be made as shown in Fig. 7-5, if the space is available and some digging is permissible. The chemicals increase the conductivity of the ground in the vicinity of the grounding pipes of rods and thus reduce the losses from current flow.

Radial Grounds

The ideal form of ground is a series of conductors buried a foot or two beneath the surface, radiating like the spokes of a wheel from under the vertical part of the antenna, as shown in Fig. 7-6. Its construction is beyond most amateurs, but it is mentioned here for the benefit of those who may have the space and a plow to cut the

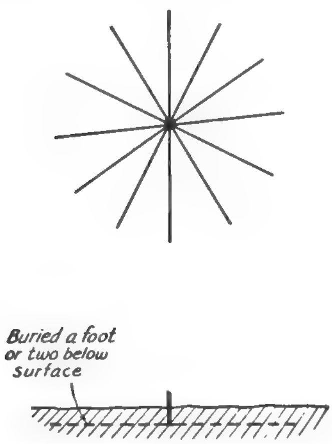


Fig. 7-6—The best ground is a radial system of buried copper strip or heavy bare copper wire.

furrows which contain the ground conductors. Such a ground system not only reduces I^2R losses at the ground connection but, provided it is made extensive enough, also greatly reduces power losses in the ground in the immediate vicinity of the antenna.

Considerations entering into the design of a radial ground system are discussed in Chapter Two. In general, it is doubtful if any real improvement over a good water-distribution-system ground will result unless at least 15 or 16 radials arranged like the spokes of a wheel, and at least ⅛ wavelength (about 70 feet) long, are used. However, less extensive radial systems may be better than driven rods in cases where no water-distribution system is available for grounding.

The Counterpoise

The counterpoise is a form of capacitive ground which is quite effective. Its use is particularly beneficial when an extensive buried system is not practicable, or when an ordinary pipe ground cannot be made to have sufficiently low resistance, as in rocky or sandy soils.

To work properly, a counterpoise must be large enough to have considerable capacitance to ground, which means that it should cover as much ground area as the location will permit. No specific dimensions have to be used, nor is the number of wires particularly critical. A good form is an approximately circular arrangement using radial wires with cross-connectors joining them at intervals, as in Fig. 7-7. There is no particular necessity for extending the radius of a circular counterpoise beyond a half wavelength, nor do the lengths of the individual wires have to bear any particular relation to the wavelength. Rather, the intention is to have the counterpoise act as a pure capacitance instead of exhibiting resonance effects. The capacitance of the counterpoise will be approximately equal to that of a capacitor consisting of two plates each of the same area as that of the counterpoise, with spacing equal to the height of the counterpoise above the ground.

The shape of the counterpoise may be made anything convenient; square or oblong arrangements are usually relatively easy to construct and will work satisfactorily. There should not be too few wires, but on the other hand separations between wires up to 10 or 15 feet will do no harm on fairly large counterpoises, and 5 to 10 feet on smaller ones. It is a good plan to join adjacent wires with jumpers at intervals about equal to the wire separation so that resonance effects will be minimized.

The height of the counterpoise is not particularly critical. It is best to construct high enough to be out of the way, which ordinarily means from 6 to 10 feet above the ground. Remember that the height of the antena is reduced by the amount of counterpoise height.

Satisfactory results have been secured with counterpoises simply lying on the ground, or with large screens of chicken wire similarly laid under the antenna. However, the best performance will be secured, as a general rule, when the counterpoise is insulated from ground. When in contact with the ground surface, the losses are likely to be higher because the counterpoise tends to act either as a poorly conducting direct ground or as a leaky-dielectric capacitor.

ANTENNAS FOR SMALL SPACE OR HEIGHT

The antennas discussed thus far have been designed to take advantage of the transmission characteristics of the 160-meter band. A certain amount of height and ground space is essential for this purpose. Many amateurs, however, do not have the facilities for the construction of even these simple forms and—particularly the city "cliff dwellers"—must simply string up a wire where some space is available.

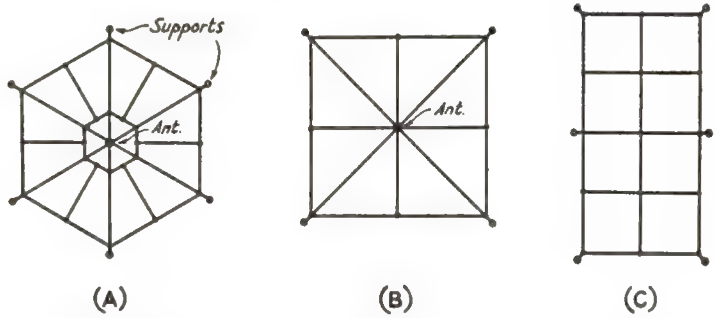


Fig. 7-7—Some suggested forms of counterpoise. Perfect symmetry is not essential, but it is desirable to extend the counterpoise as nearly as possible for the same distance in each direction from the antenna.

A vertical antenna must be quite clear of surrounding buildings, particularly those of steel construction, if good results are to be secured. If the height required for this purpose is not obtainable, then a horizontal wire must suffice. No useful purpose is served in erecting a vertical antenna between buildings which are going to absorb most of the radiated energy, or which perhaps reradiate some of the energy to make the horizontal directional pattern of the antenna poor in the most desired directions of communication.

The fundamental requirement for an antenna which cannot be "designed" for 160-meter work is that it must be resonant at the operating frequency. That is, it must accept as much power as possible from the transmitter, even though the radiation of the power must be left more or less to chance. It is desirable to get the high-current portion of the antenna well away from buildings, if this is possible. The antenna may be bent, if necessary, to fit the available space, but the bends should be made with a view to their effect on the performance of the antenna as a radiator.

To make tuning easy, it is desirable that the antenna length be a multiple of a quarter wavelength, within reasonable limits. The ground lead should be short, although, as already explained, the "length" of this lead may depend upon the grounding system. Provided the effective length of the ground lead is not too great (up to perhaps 35 or 40 feet) the system may readily be tuned to resonance with an adjustable coil and series capacitor. An antenna length of about 125 feet is the smallest recommended for working over the band, although shorter lengths can be loaded to the proper electrical length with a series inductance as shown in Fig. 7-8B.

If no space is available sufficient to allow the antenna to be installed in a straight line, it may be bent to fit. The far end may be bent, as shown in Fig. 7-8C, or even back on the antenna as in Fig. 7-8D. In the latter case at least ⅜ wavelength of the near end (the high-current part) should be unparalleled by the bent wire, since there is partial cancellation of the radiation from the folded-back part. Bends in horizontal directions may be made at several points along the wire, in cases where this is necessary, provided the angle between the bent portions is as large as possible. Try not to have less than a right-angle bend, especially in the high-current portion of the antenna.

A disadvantage if the quarter-wave "random" antenna is the fact that the high-current end, which does the most radiating, is the end brought into the station. If there is at least one quite long straight stretch available for erecting the an-

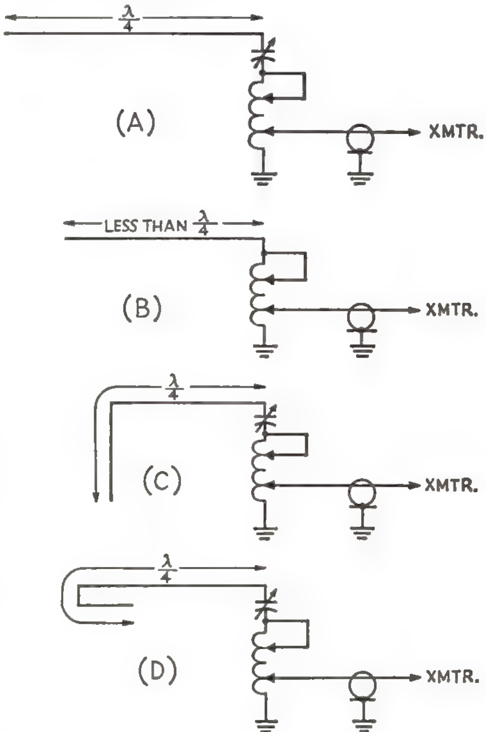


Fig. 7-8—Typical arrangements of a quarter-wave horizontal antenna, for installation where height or space is limited. A length of about 125 feet will be satisfactory.

tenna, it is a better plan to make the antenna length such that the maximum current point comes at the middle of the straight section. This means that the wire should be a quarter wave long (125 feet will be satisfactory) from the middle of the span to the far end, the necessary bends or folds to make up any excess length being made at the far end. The distance from the middle of the span to the transmitter can form the antenna length on that side or, alternatively, the wire length here may also be made a quarter wavelength, with bends or folds, to make a half-wave antenna and thus bring a voltage loop at the coupling point. The total length of the wire would be 250 feet in that case, and parallel tuning would be called for in the coupling circuit.

Antennas of this type will work quite well, especially for moderate distances at night, even though they are not capable of the type of performance to be expected from a good vertical antenna.

Alternative Coupling Methods

Other types of coupling systems may be substituted for those shown in this chapter, and Chapter Three should be consulted for design and adjustment information on this subject. In the case of vertical antennas, particularly, where the base of the antenna may be some distance from the transmitter, it may be desirable to use a transmission line to a matching circuit installed in a weather-proof box at the antenna. The feeder already described usually will be more convenient for this purpose, however, if its length fits in with the station and antenna layout.

The design and operation of such matching circuits follow the principles outlined in Chapter Three. The matching circuits are especially useful when the transmitter has a pi-network tank circuit designed for working into low-impedance loads.

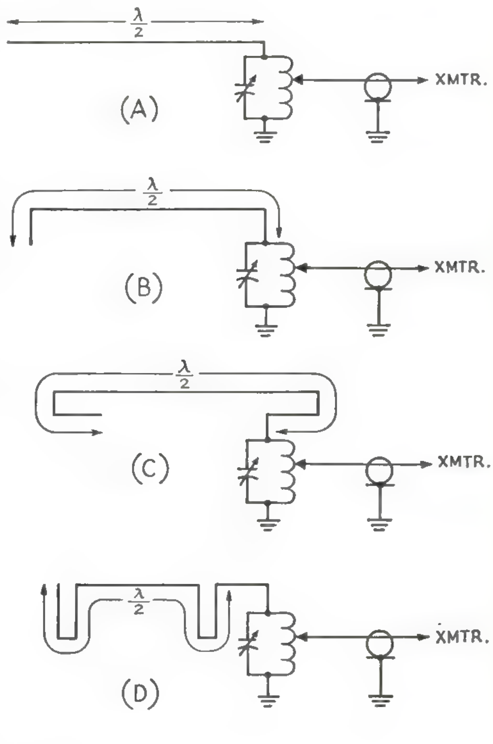


Fig. 7-9—Suggested methods of bending half-wave antennas for installation where space is limited. The points to watch out for in making bends are discussed in the text. The total wire length is about 250 feet for an antenna which can be used over the 160-meter band.

Antennas for 3.5 and 7 Mc.

Multiband antennas constructed as described in Chapter Six obviously will be useful on 3.5 and 7 Mc., and in fact the end-fed and center-fed antennas shown in Chapter Six are quite widely used for 3.5- and 7-Mc. operation. The center-fed system is better because it is inherently balanced on both bands and there is less chance for feeder radiation and r.f. feedback troubles, but either system will give a good account of itself. On these frequencies the height of the antenna is not too important, and anything over 35 feet will work well for average operation. This chapter is concerned principally with antennas designed for use on one band only.

HALF-WAVELENGTH ANTENNAS

An untuned or "flat" feedline is a logical choice on any band, because the losses are low, but it generally limits the use of the antenna to one band. Where only single-band operation is wanted the half-wave antenna fed with untuned line is one of the most popular systems on the 3.5- and 7-Mc. bands. If the antenna is a single-wire affair, its impedance is in the vicinity of 70 ohms. The most logical way to feed the antenna is with 72-ohm Twin-Lead or coaxial line. The heavy-duty Twin-Lead and the coaxial line present support problems, but these can be overcome by using a small auxiliary pole to take the weight of the line. The line should come away from the antenna at right angles, and it can be of any length.

A "folded dipole" shows an impedance of 300 ohms, and so it can be fed directly with any length of 300-ohm TV line. The line should come away from the antenna at as close to a right angle as possible. The folded dipole can be made of ordinary wire spaced by lightweight wooden or plastic spacers, 4 or 6 inches long, or a piece of 300-ohm TV line can be used for the folded dipole.

A folded dipole can be fed with a 600-ohm open-wire line with only a 2-to-1 s.w.r., but a nearly perfect match can be obtained with 600-ohm open line and a three-wire dipole.

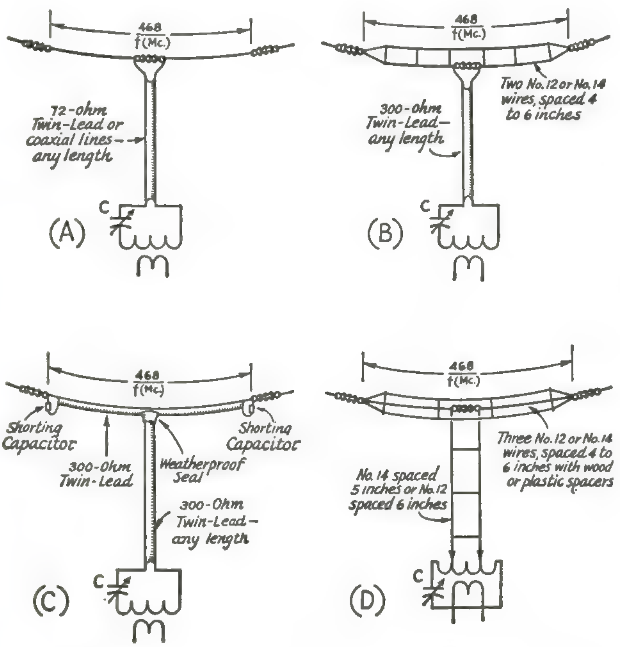
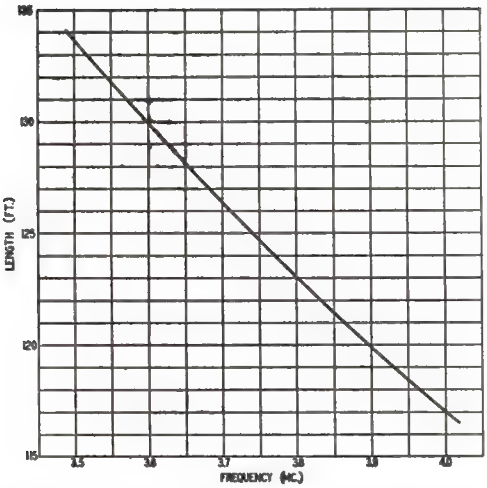


Fig. 8-1—Half-wavelength antennas for single-band operation. The multi-wire types shown in B, C and D offer a better match to the feeder over a somewhat wider range of frequencies but otherwise the performances are identical. The feeder should run away from the antenna at a right angle for as great a distance as possible. In the coupling circuits shown, tuned circuits should resonate to the operating frequency. In the series tuned circuits of A, B and C high *L* and low *C* are recommended, and in D the inductance and capacitance should be similar to the output-amplifier tank, with the feeders tapped across at least ½ the coil. The tapped-coil matching circuit shown in Chapter Six can be substituted in each case.



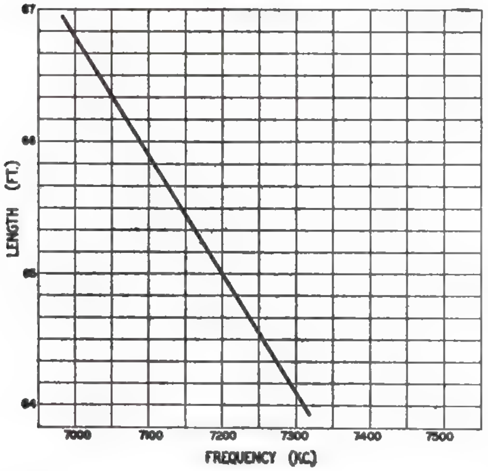


Fig. 8-2—The above charts can be used to determine the length of a half-wave antenna of wire.

The three types of half-wavelength antennas just discussed are shown in Fig. 8-1. One advantage of the two- and three-wire antennas over the single wire is that they offer a better match over a band. This is particularly important if full coverage of the 3.5-Mc. band is contemplated.

While there are many other methods of matching lines to half-wavelength antennas, the three mentioned are the most practical ones. It is possible, for example, to use a quarter-wavelength transformer of 150-ohm Twin-Lead to match a single-wire half-wavelength antenna to 300-ohm feedline. But if 300-ohm feedline is to be used, a folded dipole offers an excellent match without the necessity for a matching section.

The formula shown above each antenna in Fig. 8-1 can be used to compute the length at any frequency, or the length can be obtained directly from the charts in Fig. 8-2.

As mentioned before, the height of an antenna is not too important at these frequencies, but it should be at least 35 feet for good results. The advantages of greater height will become increasingly apparent when working DX.

Inverted-Vee Dipole

The halves of a dipole may be sloped to form an inverted V, as shown in Fig. 8-3. This has the advantages of requiring only a single high support and less horizontal space. K7GCO and others have also reported that the dipole in this form is more effective than a horizontal antenna, especially for frequencies of 7 Mc. and lower.

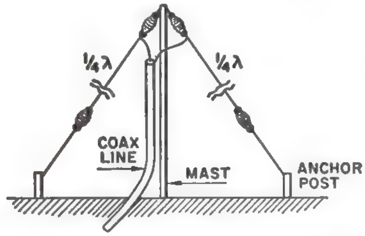


Fig. 8-3—The inverted-vee dipole. The length and apex angle should be adjusted as described in the text.

Sloping of the wires results in an increase in the resonant frequency and a decrease in feedpoint impedance and bandwidth as the angle between the two wires is decreased. Thus, for the same frequency, the length of the dipole must be increased somewhat. The angle at the apex is not critical, although it should probably be made no smaller than 90 degrees. Because of the lower impedance, a 50-ohm line should be used, and the usual procedure is to adjust the angle for lowest s.w.r. while keeping the dipole resonant by adjustment of length. Bandwidth may be increased by using multiconductor elements, such as the cage configuration.

Satisfactory two-band operation has been reported by placing 40- and 80-meter dipoles at right angles (using the same center support) and connecting the two dipoles in parallel to a single 50-ohm line, as shown in Fig. 8-4.

If no other support is available, a tree may be pressed into service.

VERTICAL ANTENNAS

Vertical antennas find some favor on 3.5 and 7 Mc., and if one has a suitable location he can

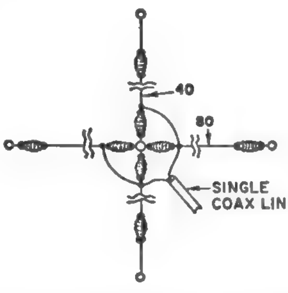


Fig. 8-4—Top view of a two-band arrangement. Dipoles for 80 and 40 meters are connected in parallel and fed with a single coaxial line.

get good results with a vertical antenna on these frequencies. There is a good chance that BCI troubles will be a little more severe with a vertical antenna, but good results in DX work can be obtained, because of the low-angle radiation from the vertical.

For 3.5-Mc. work, the vertical can be a quarter wavelength long (if one can get the height), or it can be something less than this and "top-loaded." The bottom of the antenna has only to clear the ground by inches. Probably the best construction of a quarter-wavelength vertical involves running copper or aluminum tubing alongside a wooden mast, if a metal tower is not available for use as a radiator. If a grounded tower is used, the antenna can be "shunt-fed," as shown in B of Fig. 8-5. The "Gamma" matching system described in Chapter Three may also be used. A good ground system is helpful in feeding a quarter-wavelength vertical antenna, and the ground can be either a convenient water-pipe system or a number of radial wires extending out from the base of the antenna for about a quarter wavelength.

The Ground Plane

The size of a "ground-plane" antenna makes it

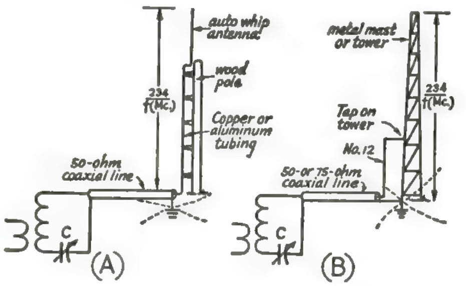


Fig. 8-5—Vertical antennas are effective for 3.5- or 7-Mc. work. The quarter-wavelength antenna shown at A is fed directly with 50-ohm coaxial line, and the resulting standing-wave ratio is usually less than 1.5 to 1, depending on the ground resistance. If a grounded antenna is used as at B, the antenna can be shunt-fed with either 50- or 75-ohm coaxial line. The tap for best match will have to be found by experiment; the line running up the side of the antenna should be spaced from 6 to 12 inches from the antenna. The length (height) of the antenna can be computed from the formula, or it can be obtained from Fig. 8-2 by using just one-half the length indicated in the chart. For example, at 3.6 Mc., the length is 120′/2 = 60′.

a little impractical for 3.5-Mc. work, but one can be used at 7 Mc. to good advantage, particularly for DX work. This type of antenna can be placed higher above ground than an ordinary vertical without decreasing the low-angle radiation. The vertical member can be a length of self-supporting tubing at the top of a short mast, and the radials can be lengths of wire used also to support the mast. The radials do not have to be exactly horizontal, as shown in Fig. 8-6.

The ground-plane antenna can be fed directly

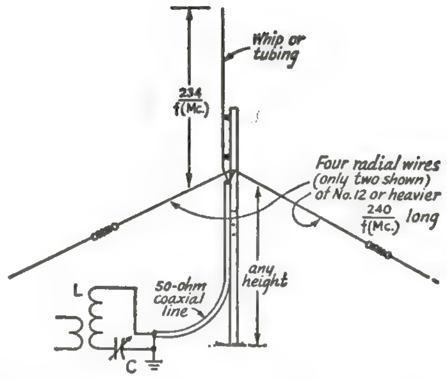


Fig. 8-6—A ground-plane antenna is effective for DX work on 7 Mc. Although its base can be any height above ground, losses in the ground underneath will be reduced by keeping the bottom of the antenna and the ground plane as high above ground as possible. Feeding the antenna directly with 50-ohm coaxial cable will result in a low standing-wave ratio. The length of the vertical radiator can be computed from the formula, or it can be obtained from Fig. 8-2 by using just one-half the length indicated in the chart. The radial wires are 2.5% longer. For example, at 7.1 Mc., the radiator is 65′ 11″/2 = 33′; the radials are 1.025 × 33 = 33′ 10″.

with 50-ohm cable, although the resulting s.w.r. on the line will not be as low as it will if the antenna is designed with a stub matching section, as described in Chapter Three. However, the additional loss caused by an s.w.r. as high as 2 to 1 will be inappreciable even in cable runs of several hundred feet when the frequency is as low as 7 Mc.

Another vertical or semi-vertical antenna that will work well on one band (either 3.5 or 7 Mc.) is shown in Fig. 8-8. A quarter-wavelength three-wire dipole of this type offers a good match for 300-ohm line over a fairly wide frequency range. The wires should be spaced from about 12 to 15 inches either side of the driven center conductor, and wood spacers boiled in paraffin are quite adequate for the job. The ground should be a good one made to a water pipe near the base of the antenna, or a counterpoise or ground-radial system can be used. It is not important that the three-wire dipole be exactly vertical—part of the far end can run horizontally or nearly so, or the entire antenna can slope as much as 45°.

If the length of the three-wire dipole is made 55 feet, the antenna will work fairly well on both 3.9 and 14 Mc.

If the aluminum tubing is available, a 3-conductor vertical antenna can be made that will be quite efficient over the entire 7-Mc. band. A pole made of two 2 × 4s fastened together to form a structure having a T-shaped cross-section, with crossarms, can be used to support the three sections of aluminum tubing, as shown in Fig. 8-7. In this instance 1-inch diameter 24ST aluminum tubing was used, spaced 15 inches on centers. The top was bolted together with a 1-inch-wide

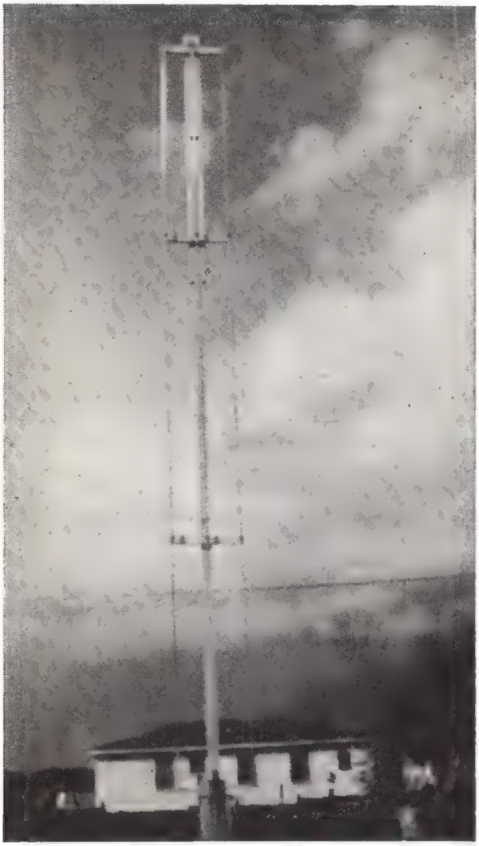


Fig. 8-7—A 3-conductor 7-Mc. vertical made of 1-inch aluminum tubing and supported by crossarms on a pole made from 2 × 4s. *(W6ECJ)*

strip of aluminum. At least four radials should be buried in the ground under the antenna, 16 radials are not an elaboration if a reasonably good ground system is desired (see Chapter Two). Unlike the 3-conductor vertical made of wire, one of aluminum tubing will require shortening the element to bring it to resonance. In the antenna pictured, a length of 30 feet 4 inches resonated at 7.1 Mc.

After all tests and measurements are concluded, the 300-ohm Twin-Lead feeder can be inserted in a length of garden hose and buried in the ground between the base of the antenna and the lead-in to the shack.

Two-Band Coax-Fed Antenna

Fig. 8-9 shows a compact antenna for 3.5 and 7 Mc. that may be fed with a single coaxial line. Essentially, the antenna is a 33-foot length of 50-ohm coaxial cable suspended vertically from a horizontal wire over a ground radial system. At the upper end of the coax cable the inner conductor is soldered to the horizontal wire as shown. The braid of the coax is removed for a distance of 6 inches from the soldered connection, leaving the polyethylene insulation on the inner conductor. Tie a nylon cord from the body of the coax cable to the horizontal wire to relieve the inner conductor from having to support the weight of the cable, and tape this 6-inch gap with plastic tape for waterproofing. Insert a large needle at the selected *A* distance from the top end of the braid, thus shorting the inner conductor to the braid at this point.

At the lower end of the coax cable, remove approximately 3 inches of vinyl, spread back the braid, and snip off the exposed polyethylene-covered inner conductor. Twist the free braid into a "rope" and fasten it to the upper eye of the base strain insulator. This is the feed point of the antenna. Wire the lower eye of the insulator to the ground stake. This lower eye is also the junction of the eight ground radials and the braid of your feeder coax.

The antenna operates as a base-driven quarter-wave vertical on 40 meters. The r.f. current flows on the outside surface of the coax braid. At the cut edge of the top it sees a high impedance as a result of the shorted quarter-wave transformer formed by the inner surface of the braid and the inner conductor, this *A* length being an electrical quarter wave in the cable. The high impedance at this point effectively decouples the flat-top wire and virtually takes it out of the picture during 40-meter operation. By adjusting the location of the needle, the point of best v.s.w.r. can be placed anywhere in the band.

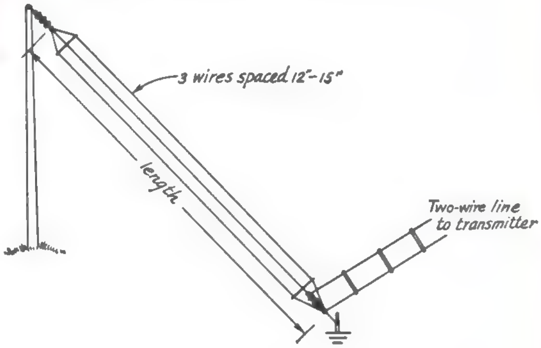


Fig. 8-8—A three-wire quarter-wave dipole for 3.5 or 7 Mc. offers a good match for 300-ohm line. The antenna does not have to run exactly vertical, but a fairly good ground is important. The length is calculated from 234 ÷ freq. (Mc.), but, because the antenna is broad, this figure is not critical.

On 80 meters the antenna operates as a top-loaded ⅛-wave vertical. The same *A* length is now much less than a quarter wavelength and is electrically equivalent to a series inductance. With this series inductance only 40 feet of single-wire flat-top is needed for operation on 80 meters.

Fine adjustments for best v.s.w.r. in the 80-meter band can be made by pruning the flat-top length. By virtue of the decoupling feature, this can be done without danger of effecting the 40-meter operation.

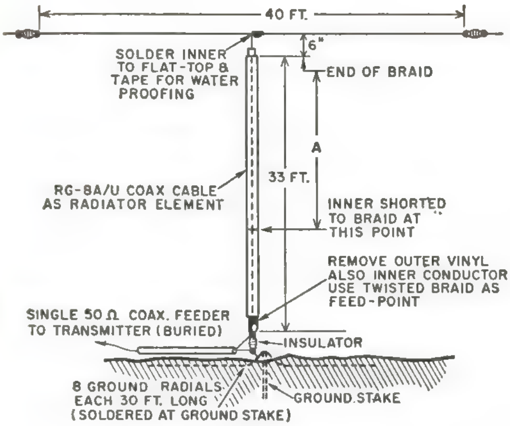


Fig. 8-9—Construction of the 40-80-meter coax-fed antenna (W2JTJ). Dimension A is an electrical quarter wavelength in the cable at the median operating frequency selected. Approximate length in feet is equal to 162 divided by the frequency in megacycles for solid-dielectric coaxial cable. Typical lengths for lowest s.w.r. are 23 feet 2 inches for 7.1 Mc. and 22 feet 10 inches for 7.2 Mc.

Conversely, fine adjustments for best v.s.w.r. in the 40-meter band do not materially affect 80 meter operation.

When the "needle short" distance is decided, a more permanent connection can be made.

The test setup of Fig. 8-10 was used for tuning up the antenna. The station receiver was used to check the frequency calibration.

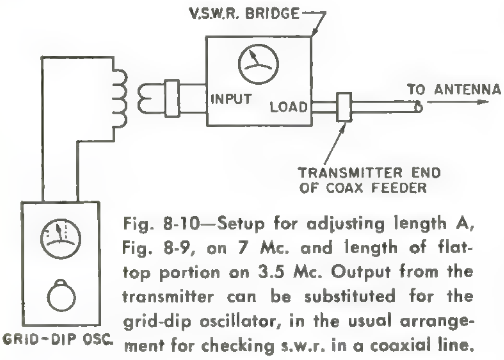


Fig. 8-10—Setup for adjusting length A, Fig. 8-9, on 7 Mc. and length of flat-top portion on 3.5 Mc. Output from the transmitter can be substituted for the grid-dip oscillator, in the usual arrangement for checking s.w.r. in a coaxial line.

PHASED ARRAYS

Phased arrays with horizontal elements can be used to advantage at 7 Mc., if they can be placed at least 40 feet above ground. Any of the usual combinations will be effective. If a bidirectional characteristic is desired, the W8JK type of array, shown at A in Fig. 8-11, is a good one. If a unidirectional characteristic is required, two elements can be mounted about 20 feet apart and provision included for tuning one of the elements as either a director or reflector, as shown in Fig. 8-11B. The parasitic element is tuned at the end of its line with a series- or parallel-tuned circuit (whichever would normally be required to couple power into the line), and the proper tuning condition can be found by using the system for receiving and listening to distant stations along the line of maximum radiation of the antenna. Tuning the feeder to the parasitic element will peak up the signal.

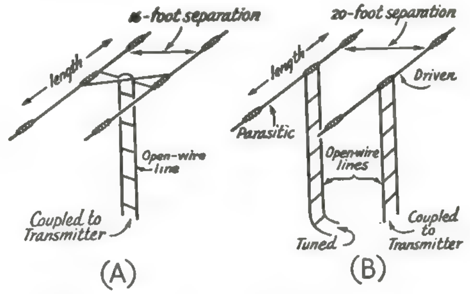


Fig. 8-11—Directional antennas for 7 Mc. To realize any advantage from these antennas, they should be at least 40 feet high. The system at A is bidirectional, and that at B is unidirectional in a direction depending upon the tuning conditions of the parasitic element. The length of the elements in either antenna should be exactly the same, but any length from 60 to 150 feet can be used. If the length of the antenna at A is between 60 and 80 feet, the antenna will be bidirectional along the same line on both 7 and 14 Mc. The system at B can be made to work on 7 and 14 Mc. in the same way, by keeping the length between 60 and 80 feet.

Antennas for 14, 21 and 28 Mc.

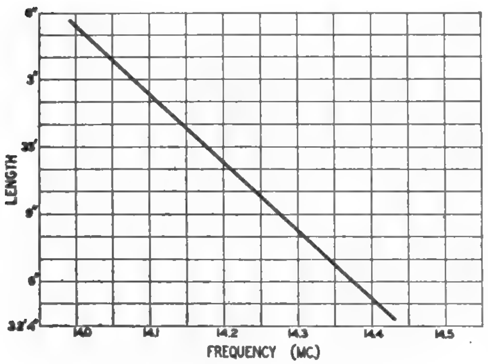
The antenna systems discussed in Chapter Six can, of course, be used on 14, 21 and 28 Mc. with good results. The half-wave antenna for 3.5 Mc., fed with tuned feeders, becomes a multi-wavelength antenna at these higher frequencies, and the directional characteristics become a little more apparent than at the lower frequencies. Similarly, a 7-Mc. half-wave antenna using tuned feeders likewise can be used on the harmonically-related higher-frequency bands. Other types of antennas that are suitable are also discussed in Chapter Six.

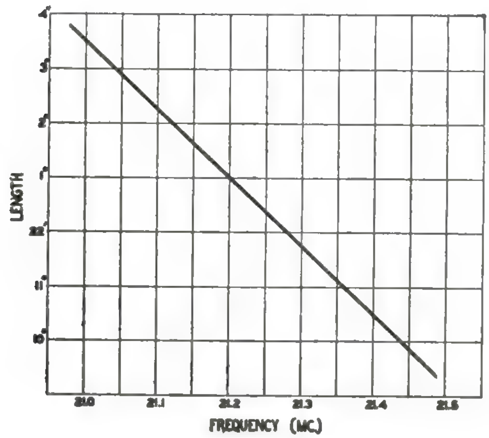
Half-Wave Dipoles

The half-wave dipole fed with a matched transmission line also is often used on the 14, 21 and 28 Mc. bands. Like its low-frequency counterpart described in Chapter Eight, it is ordinarily useful only on the band for which it is designed. Suitable lengths for wire antennas are given in Fig. 9-1, and Fig. 3-50 should be referred to if feeder resonances are to be avoided. Vertical antennas (and the ground plane in particular) can be used at these frequencies and will give good low-angle radiation. However, with a vertical receiving antenna man-made noise pickup is likely to be greater than with a horizontal antenna.

The directional pattern of a half-wavelength horizontal antenna becomes apparent at these frequencies, and it is not unwise to provide two half-wave horizontal antennas for these bands at right angles to each other, with a suitable switching arrangement to permit using one antenna or the other, depending upon the direction of the desired signal. Better still is to use a single half-wave antenna that can be rotated at least 135°.

In the case of either the switchable antennas or the rotatable half-wave, it is highly desirable to use a carefully balanced feed line, to insure against unwanted pickup of signals on the line and consequent loss of directional effects. For the same reason, it is advisable to use a balun at the antenna when a coax transmission line is used. Suitable forms of baluns are shown in Chapter Three.





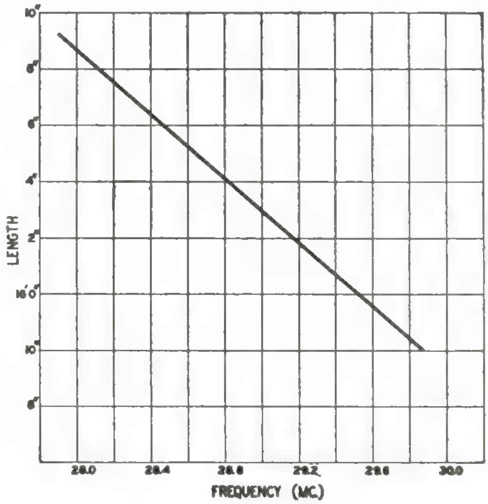


Fig. 9-1—Charts for determining the length of a half-wave wire antenna at 14, 21 and 28 Mc., based on 468/f (Mc.). Constructional details are shown in Fig. 8-1.

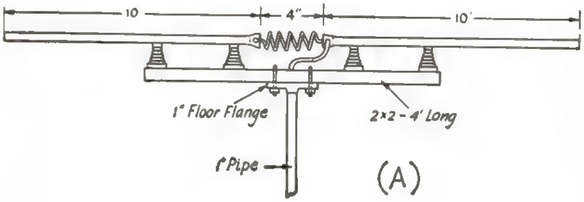
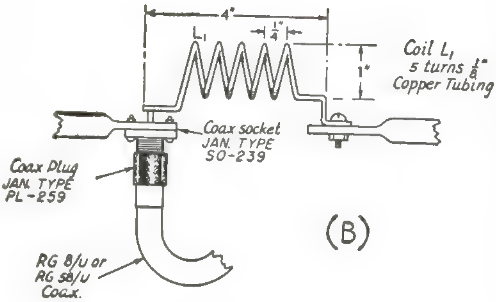


Fig. 9-2—(A) Diagram of the 21-Mc. rotatable antenna and mounting. The U bolts that hold the 2 by 2 to the floor flange are standard 2-inch TV mast type bolts. (B) A more detailed drawing of the coil and coax-fitting mountings. The ¼-inch spacing between turns is not critical, and they can vary as much as 1/16 inch without any apparent harm to the match.



Rotatable Dipole

An example of an easily-constructed rotatable half-wavelength antenna for 21 Mc. is shown in Figs. 9-2 and 9-3. The antenna is made from two pieces of ½-inch diameter electrical thin-wall steel tubing or conduit. This tubing is readily available at any electric supply shop and sells for approximately seven cents a foot. It comes in 10-foot lengths and, while 20 feet is slightly short for a half-wave antenna on 21 Mc., this length when fed at the center exhibits a resistance of approximately 50 ohms when the antenna is tuned to resonance by means of a small loading coil. Thus it offers a good match to 52-ohm coaxial cable. If aluminum tubing is available it can be used in place of the conduit, and the antenna will be lighter in weight.

The two pieces of tubing are supported by four stand off insulators on a four-foot-long 2 by 2. The coax fitting for the feed line is mounted on the end of one of the lengths of tubing. A mounting point can be made by flattening the end of the tubing for a length of about 1½ inches. The tubing can be flattened by squeezing it in a vise or by laying the end of the tubing on a hard surface and then hammering it flat. This will provide enough space to accommodate the coax fitting (Amphenol type 83-1R). A ⅝-inch hole will be needed in the flat section to clear the shell of the coax fitting.

The loading coil, L_1, is made from ⅛-inch diameter copper tubing. It consists of 5 turns spaced ¼ inch apart and is 1 inch inside diameter. The coil is connected between the inner conductor pin on the coax fitting and the other half of the antenna, Fig. 9-2B. In order to secure a good connection at the coax fitting, the coil lead should be wound around the inner-conductor pin and soldered. The other end of the coil can be connected with a screw and nut.

The antenna is mounted on a 1-inch floor flange and held in place by two 2-inch bolts. The floor flange is connected to a 12-foot length of 1-inch pipe which serves as a mast. TV antenna wall mounts can be used to support the mast.

In the installation shown in Fig. 9-3, 19-inch wall mounts were used in order to clear the eaves of the house. A 2-inch long piece of 1¼-inch pipe was used as a sleeve, and it was clamped in the *U* bolt on the bottom wall mount. A ¼-inch hole was drilled through the mast pipe approximately 6 inches from the bottom. Next, a 1½-inch bolt was slipped through the hole and the mast was then mounted in the sleeve on the bottom wall mount. The bolt acted as a bearing point against the top of the sleeve. Another ¼-inch hole was drilled through the mast about three feet above the bottom wall mount. A piece of ¼-inch metal rod, six inches long, was forced through the hole so that the rod projected on

Fig. 9-3—Over-all view of the antenna and mounting. The antenna is mounted against the side of the house, using TV wall mounts for holding the mast. The feed line comes out of the bottom of the mast and through the wall into the shack.

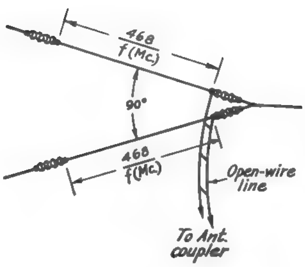


Fig. 9-4—A simple nondirectional antenna for high-frequency work. It will work satisfactorily from half the design frequency to anything higher. Both wires should be horizontal. Designed for 14.2 Mc., the length of each leg is 33 feet.

each side of the mast. To turn the mast, a piece of rope was attached to each end of the rod and the rope was brought into the shack, so that the antenna could be rotated by hand. Obviously, one could spend more money for a "de luxe" version and use a TV antenna rotator and mast.

The same type of construction can be used for 14- and 28-Mc. dipoles, with suitable modification because of the different antenna dimensions required. For 14 Mc. the antenna should be 15 feet long on each side (30 feet total) and for 28 Mc. it should be 7 feet 4 inches on a side (14 feet 8 inches total). The loading coils in each case should be adjusted in inductance to resonate the antenna at the center of the band, for general work throughout the band, or else at the most-used frequency. Approximate dimensions for L_1 for 14 Mc. are 5 turns of the same diameter and turn spacing and for 28 Mc. 3 turns. The turns can be spread apart or squeezed together to resonate the antenna. Resonance will be indicated by the lowest s.w.r., as measured by a bridge of the type described in Chapter Three, at the transmitter end of the 52-ohm coax feeder.

Nondirectional Horizontal Antenna

If one wishes to avoid the directional effects from 14 to 30 Mc., the antenna shown in Fig. 9-4 can be used. Both wires are horizontal and at right angles to each other. Since it is a balanced antenna, it will work on the three bands. If designed for 14 Mc. (each leg 33 feet long), it will be practically nondirectional on 14, 21 and 28 Mc., and it will show only slight directional effects on 7 Mc., along the bisector of the 90-degree angle.

Long Wires

When using a multiband antenna at the higher frequencies (one of the antennas described in Chapter Six), it will pay to lay out its direction to take advantage if the major lobes that develop at the higher frequencies (see Chapter Two), if this is at all possible. Further, if the antenna is not to run horizontal, it should be kept in mind that the major useful radiation from a long sloping wire is in the general direction of the downward slope although, of course, there will be useful radiation at other angles as well.

A center-fed multiband antenna will take on the directional characteristics of one-half its total length. For example, on the band where it represents two full-wave antennas in phase (over-all length of two wavelengths) it has the characteristics of a full-wave antenna but with somewhat sharper lobes. On the other hand, a multiband antenna fed at one end has the characteristic of an antenna of the total length. For example, the two-wavelength antenna fed at one and has the radiation pattern of a two-wavelength antenna.

Simple Ground-Plane Antenna

A single-band ground-plane antenna can be easily constructed from the inexpensive materials. Fig. 9-5 is an example of a 14-Mc. antenna of this type; for 21 Mc. all dimensions may be multiplied by ⅔, and for 28 Mc. the dimensions may be cut to one-half.

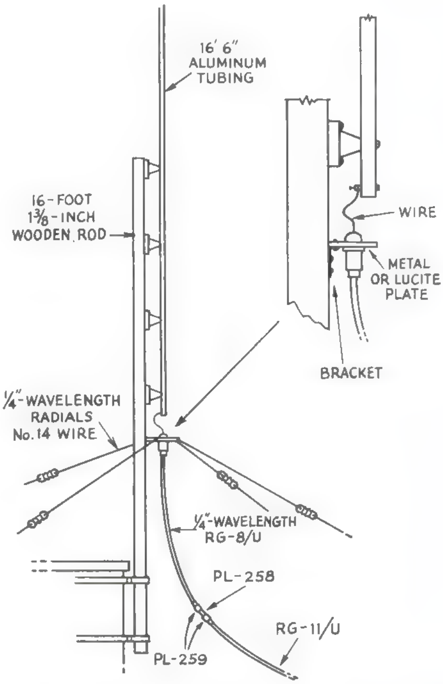


Fig. 9-5—14-Mc. ground-plane antenna (K2IKZ) using a section of 52-ohm cable for matching the antenna to 75-ohm transmission line. The wooden mast can be supported in any convenient fashion; the method shown here uses a chimney mounting with TV-type brackets.

This antenna uses a quarter-wave "Q" section to improve the match between the feed-point impedance and the 75-ohm coaxial transmission line. With horizontal radials (there should be at least four radials) the feed-point resistance is in the neighborhood of 30 ohms. This is stepped up to approximately 90 ohms through the quarter-wave section of 52-ohm cable, so that the mismatch to 75-ohm cable is small. The 75-ohm cable may be any convenient length.

If the radial wires are slanted downward the feed-point impedance increases. If the radials are sloped at about a 45-degree angle the antenna resistance will be a reasonably good match to 52-ohm cable, so cable of this characteristic impedance can be connected directly to the antenna feed point. The sloping radials obviously can be used as a part of the guying system for the mast which supports the vertical radiator.

Multiband Ground Plane

The principle employed in the multiple-dipole antennas described in Chapter 6 may be applied to the ground-plane configuration. The problem of supporting longer radiators usually limits its practical use to the 14-Mc. band and higher. A simple form of construction is shown in Fig. 9-6. A 16-ft. length of 2×2 lumber serves as the support for quarter-wave radiators for 10, 15 and 20 meters, one radiator being mounted on each of three sides of the support. Lengths of 300-ohm ribbon line, with the two conductors connected in parallel, are used for the radiators in this instance, although any single conductor would work equally well. The elements are supported on TV standoff insulators. The 20-meter radiator is slightly longer than the support, so its top end is supported as shown in the sketch.

The ground-plane radials may be made of 4-conductor TV rotator-control ribbon, cutting two of the four conductors for 20 meters, one conductor for 15 meters, and one conductor for 10 meters. There should be four or more of these radials fanning out from the base of the radiator support.

The bottom ends of all radiators should be connected together and to the inner conductor of a 50-ohm coaxial feed line. The inner ends of all radials should be similarly joined and connected to the braid of the coax.

The 2 × 2 support can be clamped to the top of a mast with U bolts. As an alternative, the 2 × 2 may be provided with a hook about two thirds of the way toward the top, as shown in the sketch, and the assembly supported in a tree by maneuvering the hook over a small branch.

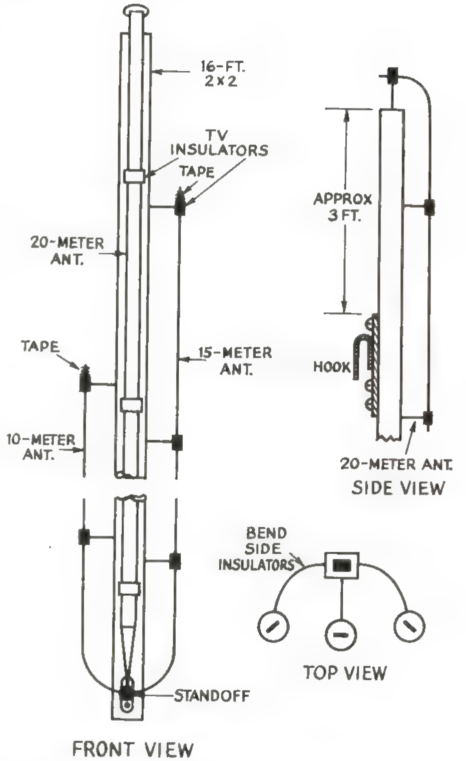


Fig. 9-6—Sketch showing the construction of a simple three-band ground-plane antenna. Radiator and radial lengths should be 16 ft. 6 inches for 14 Mc., 11 ft. 2 inches for 21 Mc., and 8 ft. 3 inches for 28 Mc.

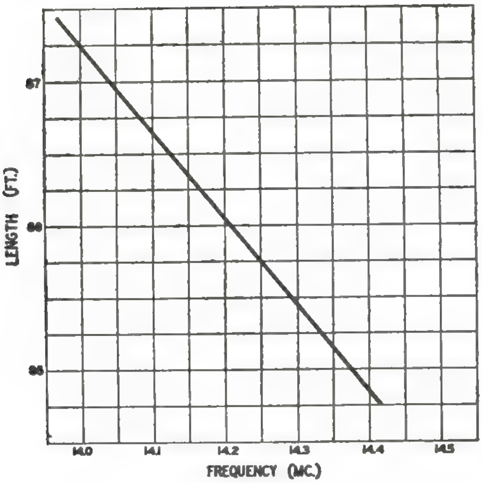
DIRECTIONAL ANTENNAS

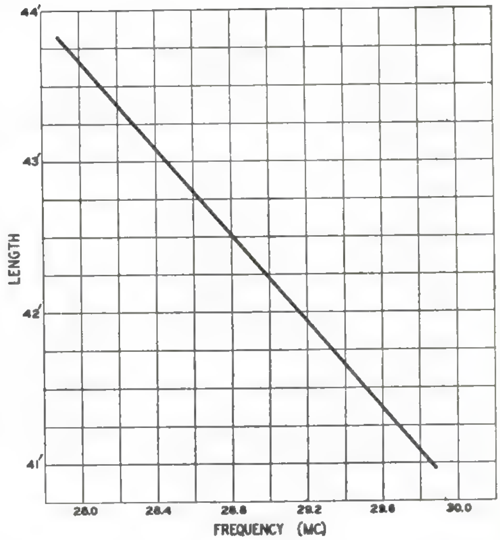
Because a wavelength is smaller physically at these frequencies than at the lower ones, relatively larger antennas (in terms of wavelengths) can be built, to take advantage of the gain and directive effects of the more complicated types of antennas. A 7-Mc. half-wavelength antenna, fed at the center with tuned line, becomes "two half waves in phase" at 14 Mc., with a gain of 1.8 db. over a 14-Mc. half-wavelength antenna. However, by making the antenna only a little longer it becomes an "extended double Zepp" and has a gain of about 3 db. The proper length of an extended double Zepp can be obtained from Fig. 9-7, and its physical construction is the same as for the center-fed antenna of Fig. 6-4. On the design frequency, and at half this frequency, the maximum radiation is at right angles to the wire. At higher frequencies, the pattern tends toward the "*X*" shapes obtained from 1- and 2-wavelength antennas, as mentioned earlier.

As one goes to the higher-frequency bands, it becomes increasingly important to confine the radiation to low angles above ground, for maximum signal strength on both transmitting and receiving. Since the vertical pattern of a simple antenna depends upon its height above ground, it would be wise to choose a height that confines the radiation to the lower angles. This is difficult to do, however, because the exact height above electrical ground is hard to determine. The solution is to use an antenna system that is less dependent upon the electrical ground level.

The ground-plane antenna will have the same pattern as a half-wavelength vertical with its center at the same height as the base of the ground plane.

Simple end-fire or broadside arrays using horizontal elements will increase the low-angle radiation. They are also fairly easy types of antennas to build. The end-fire (W8JK beam) is generally easier to install because it doesn't require as much height to be effective, but its *Q* is higher and the system is somewhat harder to feed. The feed problem can be made easier by using multiwire elements.





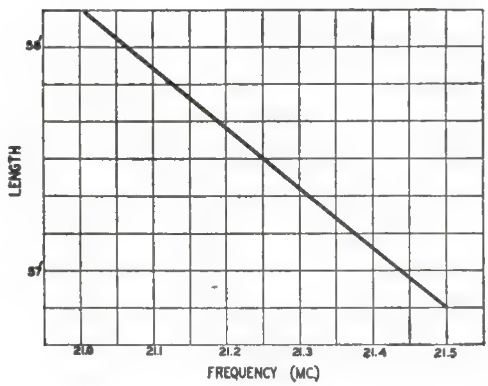
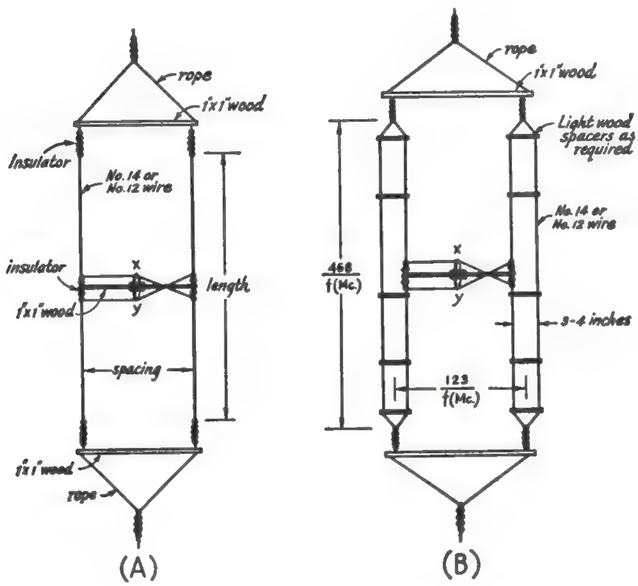


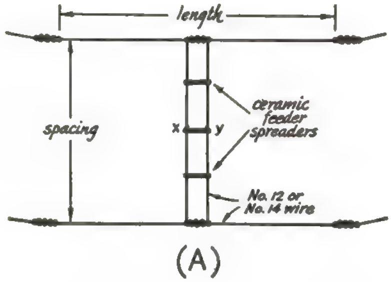
Fig. 9-7—Charts for determining the over-all length of the flat-top portion of an extended double-Zepp antenna. The construction is the same as shown in Fig. 6-4.

Fig. 9-8—Two simple end-fire arrays for 14, 21 and 28 Mc. These arrays are called "W8JK" antennas, and are generally used horizontally, thus requiring two sup- p.. .. but they can be hung vertically and rotated if desired. That shown at A will be useful on the three bands if the length is made 33 to 40 feet, with a spacing of 8½ feet.

The antenna shown at B is good for only one band, but it is easier to feed. Element lengths can be found from Fig. 9-1.

Either type of antenna is fed at points x and y with an open-wire line. The standing-wave ratio will be lower with B.





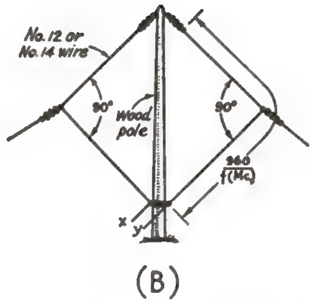


Fig. 9-9—Two popular types of broadside arrays. The "lazy H" shown at A will work on 14, 21 and 28 Mc. if the length is made 33 to 40 feet, with a spacing of 16 to 24 feet. The larger dimensions give slightly more gain on all bands. The bottom wire should be at least 15 feet above ground for three-band operation, but better results will be obtained if the bottom wire is 30 or 35 feet above ground.

The "bisquare" antenna shown at B is a useful antenna for the 28-Mc. band. It has the advantage that only one support is required. The bottom of the antenna should be at least 3 feet above the ground, and preferably about 10 or 12 feet above electrical ground. The bisquare will show some end-fire directivity at half frequency, with vertical polarization. The total length of wire on one side is 45 feet 3 inches at 21.2 Mc., and 33 feet at 29 Mc.

Either antenna should be fed at points x and y with a tuned line.

Practical types of simple end-fire and broadside arrays are shown in Figs. 9-8 and 9-9. The antenna of Fig. 9-8A shows the general form and constructional details of the "W8JK" beam. Although it is generally strung between two supports in a horizontal position, it can be hung vertically from a single support. It is useful for the amateur with little room because it can be used on the three bands, if tuned feeders are connected to points x and y. The dimensions are not critical (see caption with the sketch), but the lengths running away from points x and y should all be equal.

The folded-dipole version in Fig. 9-8 is a one-band affair, but has the advantage that the feed point (x and y) offers a more normal impedance value and the standing-wave ratio will be lower than in the first case. Open-wire line is recommended for use with either system. The longer-length elements will give more gain than the shortest recommended lengths.

The antennas shown in Fig. 9-9 are "broadside" arrays. The system illustrated in A is called the "lazy H." For best results on 14 through 28 Mc. the bottom wire should be 30 or 35 feet above ground, but the system is useful when the lower wire is only 15 or 20 feet high. Open-wire line should be used between the transmitter and points x and y. The larger dimensions will give more gain than the minimum, but any intermediate lengths can be used, if care is taken to make the system symmetrical about the feed point.

The "bisquare" array shown in Fig. 9-9B is a version of the "lazy H" that requires only one support instead of two, and its dimensions are such that it finds most application on 21 and 28 Mc. Open-wire line can be used between the transmitter and points x and y.

Unidirectional Arrays

Directivity not only increases the strength of the transmitted and received signals, but the reduction of response in undesired directions is helpful under QRM conditions. The beams and antennas described so far are at best bidirectional, and an improvement in signal strength and operating convenience can be obtained by making them unidirectional. One of the simplest

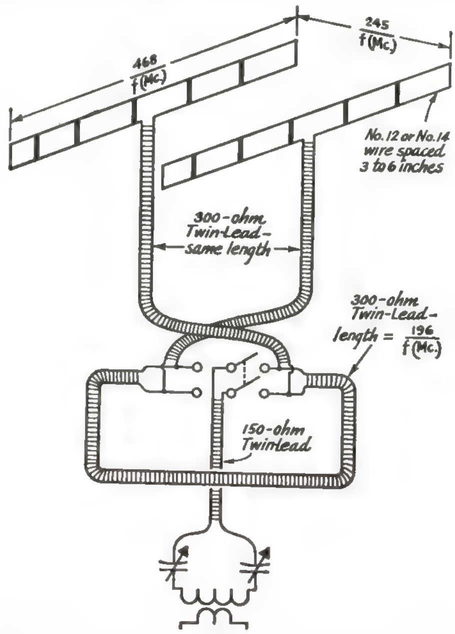


Fig. 9-10—A simple reversible fixed beam for 14, 21 or 28 Mc. Two folded dipoles (see Fig. 8-1 for constructional details) are mounted a quarter wavelength apart and are fed with 300-ohm line. The directivity is made reversible by the d.p.d.t. switch (or relay) as shown. Element lengths can be obtained from Fig. 9-1. The spacing is 17 feet 3 inches at 14.2 Mc., 11 feet 7 inches at 21.2 Mc., and 8 feet 5 inches at 29 Mc. The phasing section is 13 feet 9 inches at 14.2 Mc., 9 feet 3 inches at 21.2 Mc., and 6 feet 9 inches at 29 Mc.

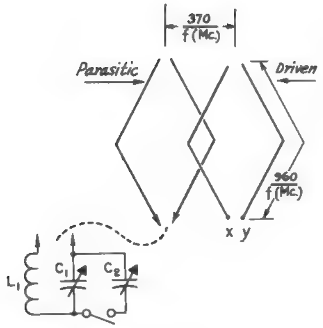


Fig. 9-11—This unidirectional beam uses two bi-square elements, one driven and one tuned as either a reflector or director. The switch is opened when the parasitic element is used as a director, and C_1 is tuned for maximum gain of the antenna in the left-hand direction. When the switch is closed, C_2 is tuned for maximum gain in the right-hand direction.

forms of unidirection beam is shown in Fig. 9-10. Two folded dipoles spaced a quarter wavelength are fed by equal lengths of 300-ohm line. An extra length of line, electrically a quarter wavelength long, is switched into one line or the other. This causes the phase difference between the two dipoles to be 90°, and the system becomes unidirectional along a line through the two antennas. Reversing the switch reverses the direction of maximum gain. A front-to-back ratio of approximately 20 db. can be obtained with this simple system. It is desirable to place it at least a half wavelength above ground.

The bisquare antenna can be used in a somewhat similar fashion by using another section and tuning it either as a director or a reflector, as shown in Fig. 9-11. The driven element is fed at x and y by a tuned line, and the switch is opened when the parasitic element is used as a director. Under these conditions, C_1 is tuned for maximum signal coming from the left-hand side (Fig. 9-11). For signals coming in the opposite direction (parasitic element used as a reflector), the switch is closed and C_2 peaked for the best signal coming from the right-hand side. Once the two capacitors have been adjusted, the switch can be controlled remotely from the operating position if desired.

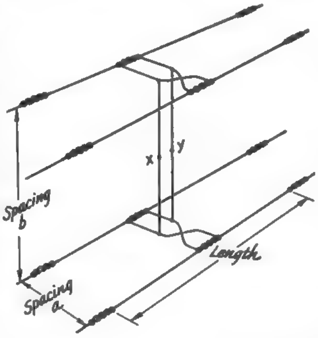


Fig. 9-12 — A bidirectional beam, made by combining the end-fire array of Fig. 9-8 with the broadside array of Fig. 9-9A. If the length is made from 33 to 40 feet, spacing a 8½ feet, and spacing b 16 to 24 feet, the beam will operate well on 14, 21 and 28 Mc. For any one band, the length can be from 0.5 to 1.2 wavelength, spacing a ⅛ wavelength, and spacing b 0.5 to 0.75 wavelength.

Combination Arrays

By combining the end-fire and broadside arrays, multiband combinations can be obtained that have improved gain and low-angle radiation. One of the best combinations is shown in Fig. 9-12. By using tuned feeders, this bidirectional beam will work well on 14, 21 and 28 Mc. Its construction is simply that of two W8JK beams (Fig. 9-8A), supported one above the other. For best results, the bottom section should be at least 30 or 35 feet above ground, but the system will work well with the bottom element only 15 or 20 feet high. For single-band operation, the beam can be fed with a flat line through a matching transformer at x and y, and the height of the bottom section should be a half wavelength. All similar sections of the system (i.e., the wires running from points x and y, the four active antenna wires etc.) should be of equal length, but the numerical value of length is unimportant within the range mentioned in Fig. 9-12.

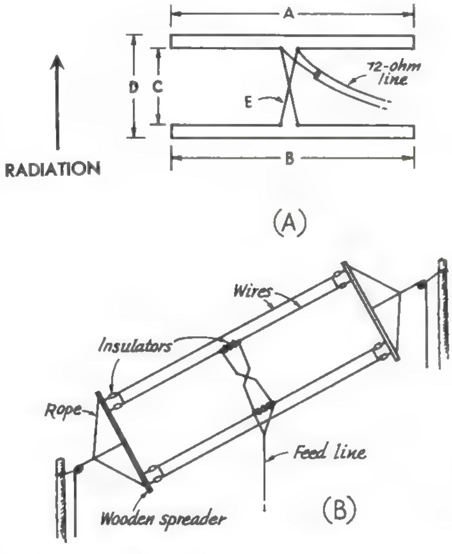


Fig. 9-13—The ZL Special two-element driven beam (A) and a possible method of construction when used as a fixed beam (B).

Dimensions are calculated from the following, where the dimensions are in feet and the frequency in Mc.:

A = 438 ÷ Mc. C = 101 ÷ Mc.
B = 447 ÷ Mc. D = 122 ÷ Mc.
E = 110 ÷ Mc. (300-ohm Twin-Lead)

The "ZL Special"

The "ZL Special" is related to the antenna shown in Fig. 9-10. It consists of two folded dipoles spaced 0.1 wavelength and driven 135 degrees out of phase. The resulting pattern is substantially unidirectional. The sketch in Fig. 9-13A shows the electrical layout and physical

Fig. 9-14—A 20-meter ZL Special mounted on a 32-foot pole and turned by a TV antenna rotator. This antenna was made of 1-inch diameter aluminum tubing; the boom is 8 feet long and the wooden crossarms are 8 feet long. Dimensions are (see Fig. 9-13): A, 30 feet 10 inches; B, 31 feet 6 inches; C, 7 feet 1 inch; D, 8 feet, 7 inches; E, 7 feet 9 inches of 300-ohm Twin-Lead. The beam is fed with RG-11/U. (*DL4SK, W4BGP*)

dimensions, and Fig. 9-13B shows one possible type of construction for a fixed beam of this type. The drive point where the feed line is attached to the antenna shows an impedance of around 70 ohms, so the antenna can be fed directly with 72-ohm coaxial line or Twin-Lead, or with 300-ohm line if a quarter-wave matching transformer of 150-ohm line is used. The estimated gain is of the order of 3 to 4 db.

The *ZL* Special can be built as a rotary beam in two general ways. One, and perhaps the simpler, is to use self-supporting elements of aluminum tubing, as shown in the example in Fig. 9-14.

The other method is to use wire elements with full-length supports, as shown in Fig. 9-15. In the latter arrangement, the beam elements and phasing line are made from 300-ohm ribbon line. Bamboo poles are used as supports for the ribbon. Such poles are obtainable from sporting-goods houses that handle fishing poles, or from rug dealers whose products are usually shipped wound around sections of bamboo. These usually come in lengths of 12 to 15 ft.

The poles are fastened to sections of 2 × 3-inch lumber, using electrical tape, pipe clamps or U bolts. The boom is a section of 2 × 4-inch lumber. The dimensions shown in Fig. 9-15 are suitable for a 15-meter antenna. For other frequencies, the boom length can be estimated from the element spacing shown in the table. The 300-ohm ribbon elements are taped to the bamboo poles.

In each of the two elements, the two conductors of the 300-ohm ribbon are connected together at the ends. One conductor of each ribbon is cut at the center for connection to the phasing line. As in Fig. 9-13, a single 180-degree twist must be made in the phasing line so that each conductor in the phasing line joins opposite sides

Table 9-I

Mc.	*A*	*B*	*C*
14.05	31′2″	31′10″	7′10″
14.25	30′9″	31′5″	7′9″
21.1	20′9″	21′2″	5′2½″
21.3	20′7″	21′0″	5′2″
28.1	15′7″	15′11″	3′10″
28.7	15′3″	15′7″	3′10″

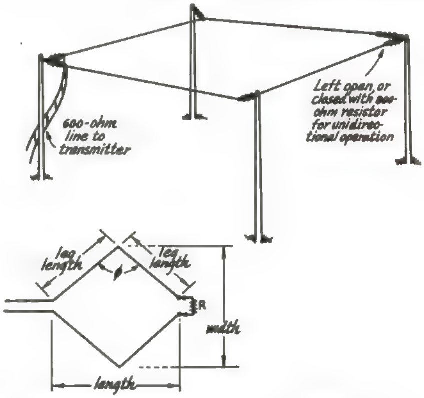


Fig. 9-16—The rhombic antenna requires four supports, but it has excellent gain and directivity over a wide frequency range. If the rhombic is terminated in a resistor, for unidirectional work, it can be fed by a 600-ohm line and a low standing-wave ratio will be obtained. An unterminated rhombic should also be fed with an open-wire 600-ohm line, but the s.w.r. will be higher. A few recommended dimensions are given below for use on 14, 21 and 28 Mc. The height should be at least 35 feet—greater heights will result in better low-angle radiation. These dimensions are compromise figures for multiband work.

Leg length (feet)	137	171	205
ϕ	107°	116°	132
Length (feet)	220	290	360
Width (feet)	163	181	195

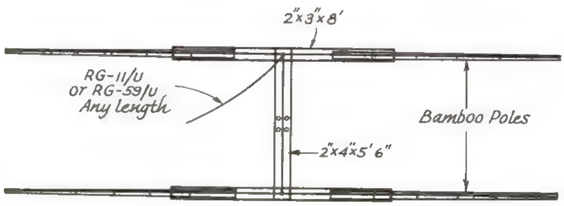


Fig. 9-15—A "ZL Special" antenna made of Twin-Lead. See Table 9-I and Fig. 9-13A for dimensions.

of the elements. Radiation is in the direction of the side to which the feed line is attached.

Vs AND RHOMBICS

On 14, 21 and 28 Mc., the long-wire beams (*V* and rhombics) are quite effective, and a properly designed rhombic will give a gain that is not likely to be equaled by any other type of antenna. The pattern of the antenna radiation becomes rather sharp, though, and hence the really high gain is effective over only a relatively narrow angle of perhaps 20 or 25 degrees. However, minor lobes of considerable amplitude help to fill in some of the other directions.

The design of Vs and rhombics is treated in Chapter Five, but it can be repeated here that the larger and higher the antenna is made, the higher will be the gain and the lower will be the wave angle. If a rhombic is terminated to give a unidirectional characteristic, one should be sure before terminating the beam that he knows over which of the two possible paths the signals generally travel. An 800-ohm noninductive resistor, rated at one-half the transmitter output power or more, should be used.

Unfortunately, the excellent results obtained with the long-wire beams do not maintain as the antenna is made smaller and smaller (for the same frequency), and so it is wiser to use some other form of beam if space is limited. The minimum rhombic dimensions shown in Fig. 9-16 are about as small as it is advisable to go. However, the investment in parts and labor in even the smallest practical rhombic is a good one, since the antenna will put out a good signal on bands from 3.5 to 30 Mc., even though the design center is one particular band and the directivity is optimum only for that frequency.

If room is available, a very effective system can be made by using a number of *V* beams with their vertices at a common support (which is generally made higher than the outlying supports, so that the antennas will slope), and running feeder lines from each antenna element. Then any V beam in the system can be selected by switching. Such an antenna is shown in Fig. 9-17, for four and seven wires. Other suitable combinations are five 5-wavelength elements arranged radially at 45 degrees, and nine 6-wavelength elements arranged radially at 40 degrees. If the elements are designed for 21 Mc., the antennas will work well from 14 to 30 Mc., and will of course have useful radiation (but not as much directivity) on the lower frequencies. The higher the wires can be strung, the greater will be the gain of the beams, and the antennas should be at least 35 feet high for any real effectiveness. The feeder wires should be spaced from 3 to 6 inches around the periphery of a circle, and all feed wires should be made the same length. The feeders in use at any time are coupled to the transmitter through a tuned coupling system, and the unused feed wires should be grounded or left floating, whichever condition shows less r.f. in the unused wires and gives the greater directional effect on receiving.

ROTARY BEAMS

On 14 Mc. and higher the mechanical problems of building an antenna which can be rotated, as a unit, to "put its best foot forward" in any desired direction are within the capabilities of the amateur operator. A few examples of antennas thus arranged to be rotated to cover all azimuthal directions have already been given in this chapter. The most used types, however, are simple directive arrays using parasitic elements. Their design and construction is a subject in itself, and therefore is given separate treatment in Chapter Twelve.

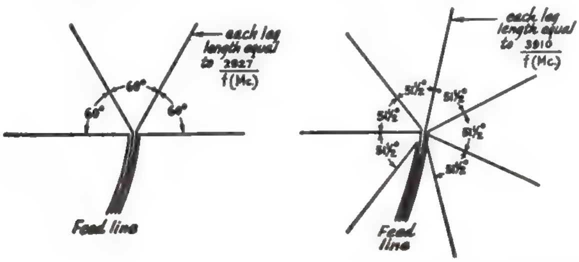


Fig. 9-17—Horizontal V beams can be arranged around a central support to give switchable directivity. The two examples are for 3- and 4-wavelength legs respectively. The system at A is not repeated for another 180° because any one V beam is bidirectional. Two legs could be omitted from the system at B with no appreciable loss of coverage.

V.H.F. and U.H.F. Antenna Systems

While the basic principles of antenna design are essentially the same for all communication frequencies, certain factors peculiar to v.h.f. and u.h.f. work call for changes in amateur antenna technique for the frequencies above 50 megacyles. Here the physical size of multielement arrays is reduced to the point where an antenna system having some gain over a simple dipole is possible in almost every location, and the more progressive stations may employ arrays having much higher gain than is possible on lower frequencies.

The importance of high-gain antennas in v.h.f. work cannot be overemphasized. The reliable working range of a station operating on 144 Mc., for instance, may be only 30 miles or so when a simple dipole antenna is used, yet this same fellow may increase his working radius to 100 miles or more by the installation of a high-gain array. The directive system introduces other advantages also. By restricting the field covered at any one position the beam antenna helps to reduce pick-up of man-made noises, and it may be instrumental in correcting interference to broadcast and television reception, by permitting communication in directions not coinciding with nearby antennas used on receivers for these services. A good antenna system is often the difference between routine operation and outstanding success in the v.h.f. field, and it is safe to say that by no other means can so large a return be obtained from a small investment as results from the erection of a high-gain antenna system.

Design Considerations

Antenna systems for the v.h.f. range are usually called upon to work over a wider frequency range than those used on lower bands; thus maximum frequency response becomes an important consideration in the design of a v.h.f. array. It may be necessary, in some instances, to include this characteristic at the expense of other attributes which might be considered desirable, such as high front-to-back ratio.

A properly matched line is of utmost importance in the proper functioning of the v.h.f. antenna system, because even with perfect matching the loss in a given line is essentially proportional to frequency (see Chapter Three). At 144 Mc., for example, the loss in a perfectly-matched line is approximately five time what it is in the same length of the same type of line at 28 Mc. Thus it may be more effective to use a high-gain array at relatively low height, rather than a simpler array at great height above ground, particularly if the antenna location is not shielded by buildings or heavy foliage in the immediate vicinity.

Height above ground is helpful, especially in cases where added height increases the distance to the visible horizon appreciably, but great height is by no means so all-important as it was once thought to be. Outstanding results have been obtained, particularly on 50 Mc., with relatively low antennas, and many 144-Mc. stations are working out successfully with arrays not more than 25 to 40 feet above ground.

The effectiveness of a v.h.f. antenna system can be increased markedly by stacking half-wave elements one above the other and feeding them in phase. Such stacking helps to lower the radiation angle, an important factor in extending v.h.f. coverage, without changing the beam width in azimuth. Several examples of stacked arrays are shown in the following pages.

The physical size of a v.h.f. array is an important factor in its performance. In receiving, the larger the area presented to an incoming signal the greater the strength of the signal at the receiver input terminals, other factors being equal. Thus an array for 432 Mc. must be the same *size* as one for 144 Mc., if an equal signal is to be received on both bands. The array for the higher frequency will require three times as many elements as the one for the lower band, if similar element configurations are used in both.

Polarization

Experience has shown that there is usually no marked difference in effective working radius with either horizontal or vertical polarization, though there are indications that horizontal may give somewhat higher signal levels over irregular terrain. The signal-to-noise ratio with horizontal systems is likely to be better, in regions where man-made noise is a serious problem. Horizontal arrays also may have some mechanical advantages. It is generally easier to build and rotate horizontal systems, particularly on the lower v.h.f. bands. Simple 2-, 3- or 4-element arrays have proven very effective in 50-Mc. work, and their use has reached the point of standardization on horizontal systems for that band.

The picture is somewhat different on 144 Mc. and higher bands. Vertical arrays are more easily constructed for these frequencies. Hundreds of

mobile stations on 144 Mc. nearly all using vertical whip antennas, usually enjoy somewhat wider coverage when the fixed stations also use vertical antennas, though the loss from cross-polarization may not be important in hilly terrain. Where the 144-Mc. band is used for emergency communication, a logical antenna setup consists of some sort of stacked (but nondirectional) collinear vertical array for the control station, and vertical whips for the portables and mobiles. Television and f.m. reception, both sensitive to interference from v.h.f. stations, use horizontal antennas, and it can be shown that interference is more troublesome when the amateur stations also use horizontal systems.

Horizontal polarization is gaining ground in amateur v.h.f. work, however, and it appears that in most areas its advantages outweigh the adverse factors. Except for emergency net operation, much of the normal 144-Mc. operation is done with horizontal antennas today. Areas on both coasts still make use of vertical, however, and anyone starting in on the v.h.f. bands should determine which polarization is in use in his locality before investing heavily in antenna installations. There is considerable polarization shift over mountainous or irregular terrain, but generally speaking, best results will be obtained when the same polarization is used at both ends of a path.

ELEMENT LENGTHS AND SPACINGS

The resonant length of an ungrounded antenna or antenna element is somewhat shorter than a half wavelength for the frequency at which it is to be used, as explained in Chapter Two. In dealing with v.h.f. antennas it is convenient to measure the length in inches. The following formula gives the resonant length of a half-wave element:

$$\textbf{Length (inches)} = \frac{5905 \times K}{\textbf{Freq. (Mc.)}}$$

where the factor K is dependent on the thickness of the antenna conductor and the frequency at which it is used. This factor is plotted in Fig. 2-4, Chapter Two, and applies to cylindrical conductors having uniform diameter throughout their length.

The length of a free-space half wavelength, together with lengths as modified by the factor K when it has the values 0.98, 0.96, and 0.94, are shown graphically for the 50-, 144-, 220-, and 420-Mc. bands in Fig. 10-1. Element spacings are based on free-space lengths, which can readily be converted from the half-wavelength values shown in the charts.

The factor K as given in Fig. 2-4 is based on theoretical considerations which necessarily do not provide for different methods of mounting or support for the element. The exact resonant length depends to some extent on constructional features of this nature. In average cases, the following formula has been found to work out well in practice:

$$\textbf{Length (inches)} = \frac{5540}{\textbf{Freq. (Mc.)}}$$

This corresponds closely to the curves for $K = 0.94$ in Fig. 10-1.

Tapered elements, in which successively smaller sizes of tubing are used either for light construction or in a collapsible element for portable use, tend to exhibit K factors associated with the smallest diameter of the taper. The exact resonant lnegth also depends on how the element is mounted; an element that is supported at its center through a boom of appreciably larger diameter than that of the element, for example, will have a slightly different resonant length than one that is insulated from its support.

Driven Elements

The length of a driven element, whether the element is a simple half-wave dipole or is the fed element in a parasitic array, is not especially critical, since slight mistuning can easily be compensated for in the adjustment of the matching system used between the element and the transmission line. However, it is generally desirable that the element length be close to resonance at the median operating frequency. The graphs of Fig. 10-1 are sufficiently accurate for this purpose.

Parasitic Elements

Optimum spacings for the elements of a Yagi array as determined by one investigator are tabulated in the section on parasitic arrays in Chapter Four. The optimum lengths for the parasitic reflector or director depend on the element spacings. The spacings are not highly critical, however, and the bandwidth is usually greater when the spacings of elements near the driven element are fairly large. Spacings of 0.2 wavelength (the half-wavelength figure given in Fig. 10-1 multiplied by 0.4) are customarily used in v.h.f. antennas. With these spacings the reflector will be approximately 5% longer than the driven element and the first director will be about 5% shorter than the driven element. If additional directors are used they should be progressively shorter than the driven element, as illustrated by the practical arrays shown later in this chapter.

When the lengths of elements in an array are given in terms of a decimal part of a wavelength, the wavelength used as a reference is usually the free-space wavelength. The free-space wavelength is equal to the value given by the top curve in each graph in Fig. 10-1 multiplied by 2. When a "half wave" is referred to in connection with antennas, the resonant length usually is meant.

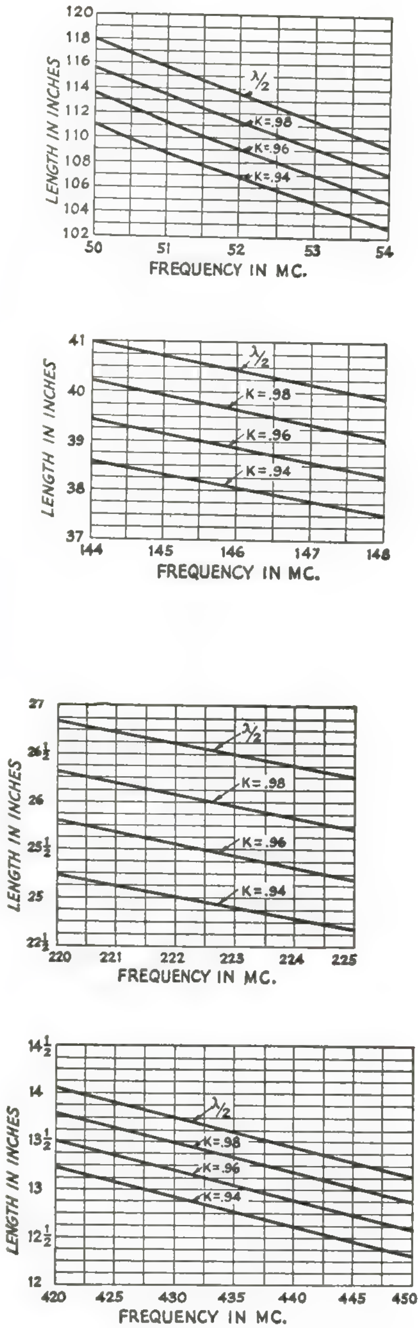


Fig. 10-1—Frequency vs. length in inches for the 50-, 144-, 220- and 420-Mc. bands, for a free-space half wavelength and resonant antenna lengths for various element length/thickness ratios (see Chapter Two). Lengths for values of K other than those given can be found by linear interpolation; e.g., the curve for $K = 0.97$ lies midway between the curves for $K = 0.96$ and 0.98, etc. To find the free-space wavelength, used in calculating spacings, multiply by 2 the length given for $\lambda/2$.

Adjustment of Element Length

When an antenna design given in this chapter is to be duplicated it is necessary to do nothing more than to cut the elements to the lengths given. If the antenna is to be centered on a frequency in a different part of the band, it will be sufficient to scale all the lengths in proportion to the change in wavelength—that is, in inverse proportion to the ratio of the desired center frequency to that for which the antenna was originally designed.

For experimental work it is desirable to have some means for continuous adjustment of element length. Usually the required adjustment range will be not greater than 10% of the half wavelength, and in the case of tubing elements a device suitable for the purpose is shown in Fig. 10-2. It consists of a short length of tubing, usually of the same stock as the element, slotted lengthwise and compressed to make a tight fit in the element end.

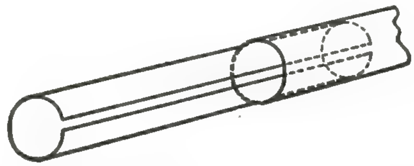


Fig. 10-2—A simple method of providing for adjustment of element lengths. The insert is made of the same size tubing as the element, but is slotted and compressed to permit insertion into the ends of the elements.

PHASING AND MATCHING SECTIONS

Transmission-line lengths for such applications as phasing lines and Q sections can be determined from the formula

$$\text{Length (inches)} = \frac{5905 \times V}{\text{Freq. (Mc.)}}$$

where V is the velocity factor of the type of line used. For open-wire lines separated by insulating spacers V is approximately 0.975. Parallel-conductor lines of self-supporting tubing have a velocity factor close to unity. The velocity factors of various types of solid-dielectric lines, both parallel-conductor and coaxial, are given in Chapter Three.

PRACTICAL ANTENNA DESIGNS

The element lengths and spacings in the antennas described in this chapter have been worked out by experiment to meet practical operating requirements. Since these requirements often are conflicting, in terms of antenna

design, compromises are necessary; for example, it it usually necessary to sacrifice a small amount of gain for the sake of reasonable bandwidth and a good front-to-back ratio. In general, this means that the element lengths and to a lesser extent the spacings (the latter are less critical than the element lengths) may not always conform to the values that investigation has shown to be optimum for maximum gain.

It can be emphasized, however, that a design that has been carefully worked out will, if accurately duplicated, give results identical with those obtained with the original antenna. If the builder has other objectives than the designer originally had, they may be achieved by individual adjustment; careful adjustment of lengths and spacings for a desired result, whether it be maximum gain, high front-to-back ratio, maximum bandwidth, or whatever, is a process that will have a great deal of appeal for the experimentally-inclined amateur.

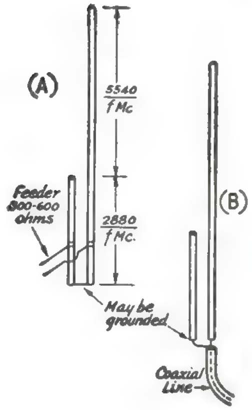


Fig. 10-3—Two versions of the "J" antenna, used in mobile applications, or in vertical arrays where parasitic elements are rotated around a fixed radiator.

The "J" is also useful in mobile applications, though a simple quarter-wave whip will usually suffice for mobile work.

TRANSMISSION LINES AND MATCHING METHODS

As mentioned at the beginning of this chapter, it is important that the standing-wave ratio on the transmission line be kept as low as possible. Otherwise line losses may become prohibitive, particularly when solid-dielectric lines are used. Lines normally employed include open-wire lines of 300 to 600 ohms impedance, usually spaced ¼ to 2 inches, polyethylene-insulated flexible parallel-conductor lines, available in 72-, 150- and 300-ohm impedances, and coaxial lines of 50 to 90 ohms.

Occasionally two coaxial lines are used side by side, with the inner conductors serving as the transmission line, and the outer conductors connected together and grounded. Such a line has twice the impedance of its individual components.

The various types of transmission lines can be matched to antenna systems in a wide variety of ways, as described in Chapter Three. The more popular methods are described below.

The "J"

Used principally as a means of feeding a stationary vertical radiator, around which parasitic elements are rotated, the "J" consists of a half-wave vertical radiator fed by a quarter-wave stub matching section, as shown at A, Fig. 10-3. The spacing between the two sides of the matching section should be two inches or less, and the point of attachment of the line will depend on the impedance of the line used. The feeder should be slid along the matching section until the point is found that gives the best operation. The bottom of the matching section may be grounded for lightning protection.

A variation of the "J" for use with coaxial-line feed is shown at B in Fig. 10-3.

The Delta or "Y" Match

Probably the simplest arrangement for feeding a dipole or parasitic array is the delta, or "Y" match. Information on figuring the dimensions of the delta may be found in Chapter Three.

The chief weakness of the delta is the likelihood of radiation from the matching section, which may interfere with the effectiveness of a multielement array. It is also somewhat unstable mechanically, and quite critical in adjustment.

The T Match

The T match, shown in practical form in Fig. 10-4, provides a means of adjustment by sliding the clips along the parallel conductors, and its rigid construction is quite suitable for rotatable arrays. It may be used with a pair of coaxial lines of any impedance, or with the various other balanced transmission lines. The position of the clips should, of course, be adjusted for minimum standing-wave ratio (see Chapter Three). The T system is particularly well suited for use in all-metal "plumbing" arrays.

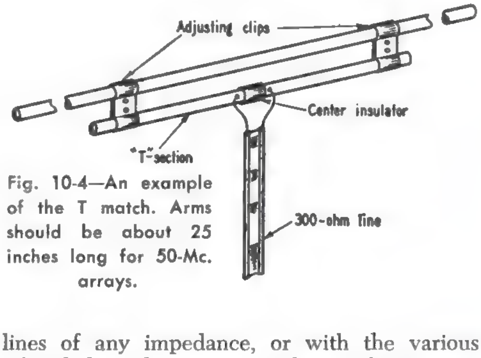


Fig. 10-4—An example of the T match. Arms should be about 25 inches long for 50-Mc. arrays.

The Gamma Match

The gamma match, also described in Chapter Three, is well adapted to feeding all-metal arrays with coaxial line, as the outer conductor may be connected to the metal boom, or to the center of the driven element. The inner conductor is tapped out from the middle of the driven element, usually through an adjustable clip. Construction can be similar to that shown for the T match.

Best operation results when a variable capacitor, *C* in Fig. 10-5, is included, to tune out the reactance of the matching section. The capacitor and the point of connection on the driven element should be adjusted for minimum standing-wave ratio. The capacitance required will be under 75 pf. for 50 Mc., or 25 pf. for 144 Mc. The r.f. voltage at this point is low, so a receiving-type variable can be used for *C*. It should be provided with a weatherproof housing, which can be of metal, grounded to the boom of the array. The length of the matching section, *D*, will be about 10 inches for 50 Mc. or 4 inches for 144 Mc.

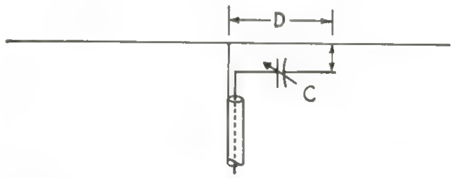


Fig. 10-5—Schematic version of the gamma match Values for C and D are given in the text.

The Q Section

As described in Chapter Three, a Q section can be used as an impedance transformer to match transmission lines to antenna systems of differing impedance. The matching section can be made of two pieces of wire, rod or tubing of suitable diameter, spaced to give the desired impedance. A table giving the impedance of lines of various dimensions may be found in Chapter Three.

Where the impedance that will be needed is not known, a Q section can be made with one of the members movable, as shown in Fig. 10-6. The spacing may then be adjusted for minimum standing-wave ratio on the transmission line.

Sections of coaxial line may be used for matching unbalanced lines to unbalanced loads, and some matching problems with balanced lines and loads can be solved with Q sections of Twin-Lead of suitable impedance. Where lines having other than air insulation are used as impedance transformers, their length should be reduced to take the propagation factor of the line into account. This will usually mean reductions to 66 and 82 per cent of a full quarter wavelength, for polyethylene-insulated coaxial and parallel lines, respectively. The exact length for a matching section can be determined experimentally, if desired, by shorting one end of the line, coupling it to a calibrated grid-dip oscillator, and then trimming the line length until the grid-dip meter shows it to be resonant at the desired frequency.

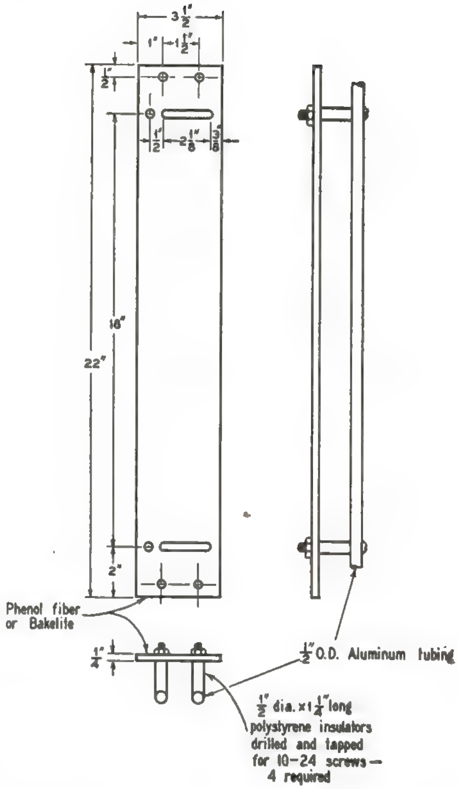


Fig. 10-6—An adjustable Q section for 2-meter arrays.

The Folded Dipole

An effective means of matching various balanced lines to the wide range of antenna impedances encountered in v.h.f. antenna work is the folded dipole (see Chapter Three), shown in it simplest form in Fig. 10-7. The simple folded dipole of Fig. 10-7 has a feed-point impedance of approximately 288 ohms. It may be fed with the popular 300-ohm line without appreciable mismatch.

The impedance at the feed point of a folded dipole may also be raised by making the diameter of the conductor used for the fed portion of the dipole smaller than the conductor used for the parallel section. Thus, in the 50-Mc. array shown in Fig. 10-17, the relatively-low center impedance is raised to a point where it may be fed directly with 300-ohm line by making the fed portion of the dipole of ¼-inch tubing, and the parallel section of 1-inch. A 3-element array of similar dimensions could be matched by sub-

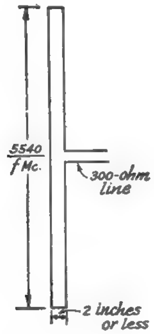


Fig. 10-7—Details of a folded-dipole for v.h.f. use.

stituting ¾-inch tubing in the unbroken section. conductor ratios and spacings for other applications may be obtained from the folded-dipole nomogram in Chapter Three.

Stub Matching

The design and adjustment of the stub matching system shown in Fig. 10-8 are described in detail in Chapter Three. For experimental work the stub may be made of tubing and the connections to it made with sliding clips. In a permanent installation a stub of open-wire line, with all connections soldered, may be more satisfactory mechanically. The transmission line may be open-wire or Twin-Lead tapped directly on the stub. Coaxial line also may be used, but should be connected to the stub through a balun.

As described in Chapter Three, the adjustment procedure is one of varying the position of the line taps and the position of the shorting taps until the lowest possible s.w.r. on the transmission line is obtained.

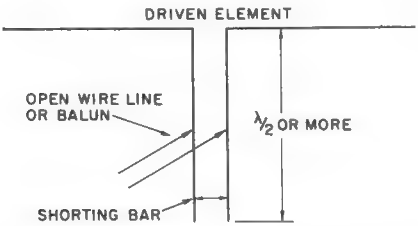


Fig. 10-8—Combination tuning and matching stub for v.h.f. arrays. The sliding shorting bar is used for tuning the driven element along with the stub itself. The transmission line (or balun, if coax line is used) is moved along the stub until the point at which the s.w.r. is closest to unity is found.

USING COAXIAL LINES

Flexible coaxial line has many desirable features. It is weatherproof, and may be buried underground, or run inside a metal mast or tower without harmful effects. However, unless it is used properly, losses may be excessive, particularly on high frequencies, when the line is more than a few wavelengths long. When coaxial line

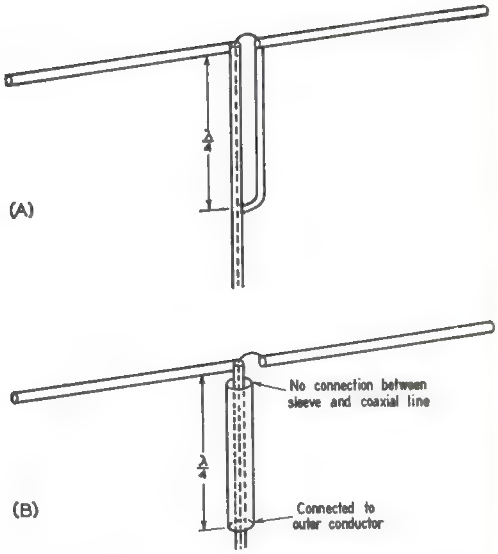


Fig. 10-9—A "bazooka" line balancer is used to feed a balanced center-fed dipole with a coaxial line. In A it takes the form of a quarter-wave section of tubing the same size as the coaxial line. In B it is a metal sleeve connected to the outer conductor at the far end from the antenna.

is used to feed a balanced load, as at the center of a driven element, some provision should be made for converting from the unbalanced line to the balanced load. Otherwise r.f. currents will flow on the outer conductor of the line, destroying its effectiveness.

One way of doing this is to install a detuning sleeve or "bazooka" at the point where the line connects to the driven element, as shown in Fig. 10-9. Both methods shown employ quarter-wave sections of line, shorted at the bottom end, presenting a high impedance to r.f. energy at the top end and preventing its flow along the outer conductor of the transmission line. The detuning section may be a piece of rod or tubing of a diameter similar to the coaxial line, as in Fig. 10-9A, or it can be a cylindrical sleeve, shorted to the outer conductor at the bottom, but insulated from it elsewhere, as in B. In either case, the length of the detuning element is a full quarter wavelength; the propagation factor of the line does *not* enter into the picture here.

Another device for feeding balanced loads with coaxial line, is called a "balun," and may take several forms. These methods also provide a 4-to-1 impedance step-up at the balanced end. A balun may be simply a folded half wavelength of coaxial line, connected as shown in Fig. 10-10.

A third method is the use of an inductively-coupled matching circuit, as shown in Fig. 10-11. The coupler may be at the transmitter, anywhere along the transmission line, or in the antenna assembly itself. It should be tuned for

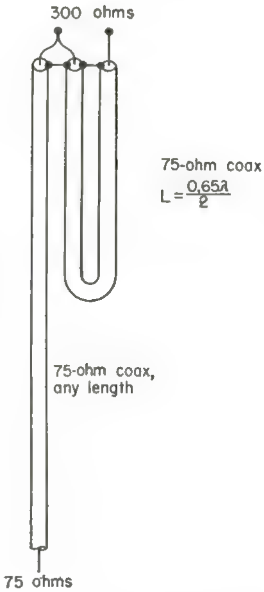


Fig. 10-10—A balun made of coaxial line may be used to feed balanced loads from unbalanced lines, and vice versa. It provides an impedance step-up of 4 to 1.

minimum standing-wave ratio on the coaxial line, and then the transmitter loading should be adjusted to the desired value.

Practical couplers of this type are shown in Fig. 10-12, one designed for 50 Mc. and the other for 144 Mc. With the taps on L_2 placed as specified in Fig. 10-11 these circuits will match balanced loads in the range 100-1600 ohms to a coaxial line (50 or 75 ohms) and so are suitable for use with 300- to 450-ohm parallel-conductor lines even when these lines are operating at a moderately-high standing-wave ratio. If a 75-ohm balanced line is connected at J_2 the taps should be moved toward the center of L_2.

As shown in Fig. 10-12, the couplers are housed in aluminum utility boxes, complete shielding being desirable. These boxes are 3 by 4 by 6 inches, and are the two-piece variety. All the components are mounted on one of the pieces. With only slight modification a standard chassis could be used, the shielding being completed by adding a bottom cover.

The two units use similar components. The main tuning capacitor, C_2, is fastened to the front wall 1¼ inches from the left side. The series capacitor, C_1, and the coaxial fitting, J_1, are 1⅛ inches up from the bottom of the rear wall and 1⅛ and 2¾ inches, respectively, from the left edge, viewing from the back. A ceramic crystal socket, J_2, is the terminal for the balanced line. It is mounted on top, one inch from the edge.

The 50-Mc. coils are cut from commercially-available stock inductors. The coupling winding, L_1, is inserted inside the tuned circuit. The plastic strips on which the coils are wound keep the two coils from shorting to each other, so no mechanical support other than that provided by the leads is needed. The leads to L_1 are brought

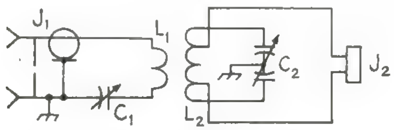


Fig. 10-11—Matching circuit for coupling balanced to unbalanced lines.

C_1—100-pf. variable for 50 Mc., 50-pf. for 144 Mc. (Hammarlund MC-100 and MC-50).
C_2—35-pf. per-section split-stator variable, 0.07-inch spacing (Hammarlund MCD-35SX). Reduce to 4 stator and 4 rotor plates in each section in 144-Mc. coupler for easier tuning; see text.
J_1—Coaxial fitting, chassis-mounting type.
J_2—Ceramic crystal socket.
L_1—50 Mc.: 4 turns No. 18 tinned, 1 inch diameter, 8 turns per inch. (Air-Dux No. 808T).
144 Mc.: 2 turns No. 14 enam., 1 inch diameter, ⅛-inch spacing. Slip over L_2 before mounting.
L_2—50 Mc.: 7 turns No. 14 tinned, 1½ inch diameter, 4 turns per inch. (Air-Dux No. 1204). Tap 1½ turns from each end.
144 Mc.: 5 turns No. 12 tinned, ½ inch diameter, ⅞ inch long. Tap 1½ turns from each end.

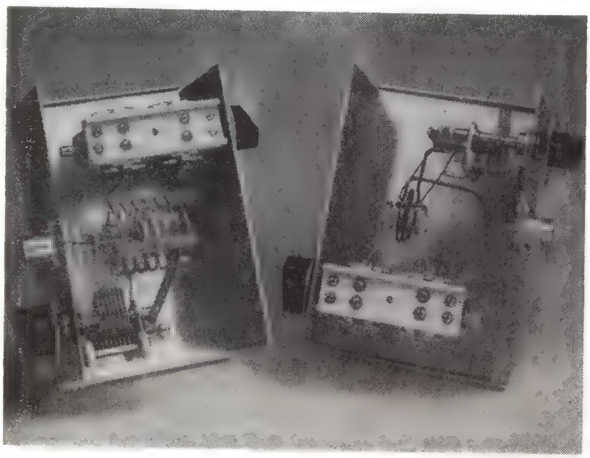


Fig. 10-12—Practical construction of matching circuits of the type shown in Fig. 10-11, for 50 Mc., left, and 144 Mc., right. Each is built on one piece of a two-piece aluminum box. These views are of the inside of the top piece.

out between the turns of L_2, and are insulated from them by two sleeves of spaghetti, one inside the other. Do not use the soft vinyl type of sleeving, as it will melt too readily if, through an accident to the antenna system, either coil should run warm. In the 144-Mc. unit the method of assembling of the coils is reversed, the tuned circuit coil, L_2, being inside the coupling coil.

The components used are adequate for fairly high power. Similar tuning capacitors are used in both couplers, but some of the plates are removed from the one in the 144-Mc. unit. This provides easier tuning, but the capacitor may be left in its original condition if desired.

The method of adjusting this type of matching circuit is covered in Chapter Three. Use of an s.w.r. bridge is highly recommended, since the proper settings of C_1 and C_2 will be those that result in the lowest possible s.w.r. in the coax line between the transmitter and the matching circuit.

Loading on the transmitter final amplifier should be adjusted at the transmitter, *after* the s.w.r. in the coax line has been brought as close as possible to 1 to 1.

Antenna Systems for 50 and 144 Mc.

A simple dipole may be used on 50 and 144 Mc. if a more pretentious antenna cannot be installed, but it is highly recommended that some form of directional array be used. While any of the designs that follow can be adapted to either band, practical considerations usually call for the use of Yagi-type arrays on 50 Mc. Yagi configurations are also employed on higher frequencies, but the small size of the elements needed for 144 Mc. and higher makes it practical to use collinear arrays, stacked Yagis, corner-reflector arrays and other more complex systems on these bands.

YAGI ARRAYS FOR 50 MC.

More gain per element probably is possible with the Yagi type of array than with any other. V.h.f. Yagi arrays usually employ wide spacing of the elements: 0.15 wavelength or more for the reflector and 0.2 wavelength or more for the directors are commonly used. Closer spacings than these tend to make the array tune too sharply to be useful across an appreciable portion of a v.h.f. band. Lack of sufficiently broad frequency response, even with wide spacing, is a problem in nearly all antenna designs for 50 Mc.

3-Element Lightweight Array

The 3-element 50-Mc. array of Fig. 10-13 weighs only 5 pounds. It uses the closest spacing that is practical for v.h.f. applications, in order to make an antenna that could be used individually or stacked in pairs without requiring a cumbersome support. The elements are half-inch aluminum tubing of 1/16-inch wall thickness, attached to the 1¼-inch dural boom with aluminum castings made for the purpose. (Dick's, RR1, Tiffin, Ohio, Type HASL.) By limiting the element spacing to 0.15 wavelength the boom is only 6 feet long. Two booms for a stacked array can thus be cut from a single 12-foot length of tubing.

The folded-diple driven element has No. 12 wire for the fed portion. The wire is mounted on ¾-inch cone standoff insulators and joined to the outer ends of the main portion by means of metal pillars and 6/32 screws and nuts. When

Fig. 10-13—Lightweight 3-element 50-Mc. array. Feeder is 52-ohm coax, with a balun for connection to the folded-dipole driven element. Balun may be coiled as shown, or taped to supporting pipe.

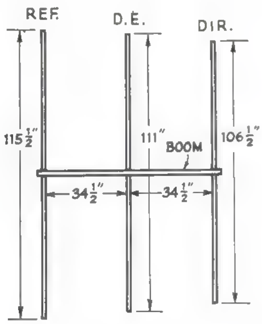


Fig. 10-14—Dimensions of the 3-element antenna shown in Fig. 10-13, for working in low-frequency section of the 50-Mc. band. The elements are ½-inch aluminum tubing. The driven element in Fig. 10-13 is a folded dipole using No. 12 wire for fed section, as described in the text.

the two halves are pulled up tightly and wrapped around the screw, solder should be sweated over the nuts and screw ends to seal the whole against weather corrosion. The same treatment should be used at each standoff. Mount a soldering lug on the ceramic cone and wrap the end of the lug around the wire and solder the whole assembly together. These joints and other portions of the array may be sprayed with clear lacquer as an additional protection.

The inner ends of the fed section are 1½ inches apart. Slip the dipole into its aluminum casting, and then drill through both element and casting with a No. 36 drill, and tap with 6/32 thread. Suitable inserts for mounting the standoffs can be made by cutting the heads off 6/32 screws. Taper the cut end of the screw slightly with a file and it will screw into the standoff readily.

Cut the dipole length according to Fig. 10-1 for the middle of the frequency range you expect to use most. The reflector and director will be approximately 4% longer and shorter, respectively. The closer spacing of the parasitic elements (0.15 wavelength; makes this deviation from the usual 5% desirable.

The folded dipole gives the single 3-element array a feed impedance of about 200 ohms at its resonant frequency. Thus it may be fed with a balun of the type shown in Fig. 10-10, using 52-ohm instead of 75-ohm coax. A gamma-matched dipole may also be used, suggested construction being as shown in Fig. 10-15. If the gamma match and 72-ohm coax are used, a balun will convert to 300-ohm balanced feed, if Twin-Lead or 300-ohm open-wire TV line feed is desired. If the dimensions are selected for optimum performance at 50.5 Mc. the array will show good performance and a fairly low standing-wave ratio over the range from 50 to 51.5 Mc.

A closeup of a mounting method for this or any other array using a round boom is shown in Fig. 10-16. Four TV-type U bolts clamp the horizontal and vertical members together. The metal plate is about 6 inches square. If ¼-inch sheet aluminum is available it may be used alone, though the photograph shows a sheet of 1/16-inch stock backed up by a piece of wood of the same size for stiffening. Tempered Masonite is preferable to wood, as it will stand up better in weather.

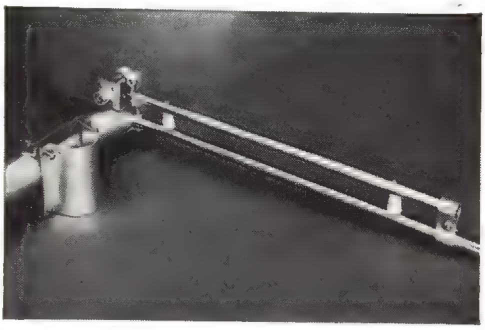


Fig. 10-15—Typical gamma match construction. The variable capacitor, 50-pf., should be mounted in an inverted plastic cup or other device to protect it from the weather. The gamma arm is about 12 inches long for 50 Mc., 5 inches for 144 Mc.

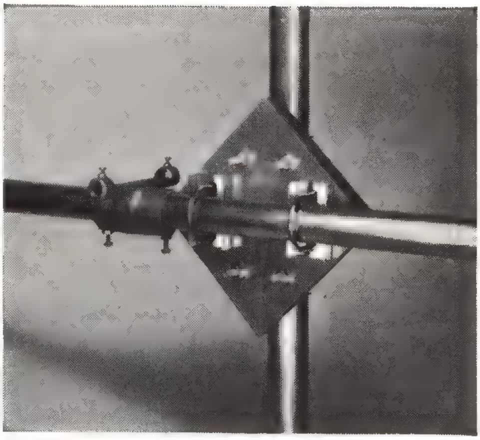


Fig. 10-16—Closeup photograph of the boom mounting for the 3-element 50-Mc. array. A sheet of aluminum 6 inches square is backed up by a piece of wood of the same size. TV-type U clamps hold the boom and vertical support together at right angles. At the left of the mounting assembly is one of the aluminum castings for holding the beam elements.

High-Performance 4-Element Array

The 4-element array of Fig. 10-17 was designed for maximum forward gain, and for direct feed with 300-ohm balanced transmission line. The parasitic elements may be any diameter from ½ to 1 inch, but the driven element should be made as shown in the sketch. The spacing between driven element and reflector, and between driven element and first director, is 0.2 wavelength. Between the first and second directors the spacing is 0.25 wavelength.

The same general arrangement may be used for a 3-element array, except that the solid por-

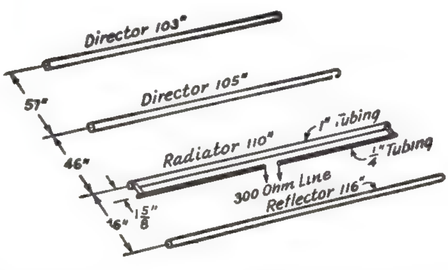


Fig. 10-17—Dimensional drawing of the 4-element 50-Mc. array. Element lengths and spacing were derived experimentally for maximum forward gain at 50.5 Mc.

tion of the dipole should be ¾-inch tubing instead of 1-inch.

With the element lengths given the array will give nearly uniform response from 50 to 51.5 Mc., and usable gain to above 52 Mc.

If a shorter boom is desired, the reflector spacing can be reduced to 0.15 wavelength and both directors spaced 0.2 wavelength, with only a slight reduction in forward gain and band width.

5-Element 50-Mc. Array

As aluminum or dural tubing is usually sold in 12-foot lengths this dimension imposes a practical limitation on the construction of a 50-Mc. beam. A 5-element array that makes optimum use of a 12-foot boom may be built according to Fig. 10-18. If the aluminum casting method of mounting elements shown for the 3-element array is employed the weight of a 5-element beam can be held to under 10 pounds.

The gamma match and coaxial line are recommended for feeding such an array. If it is desired to use 300-ohm line because of its lower losses, the gamma can be driven through a section of 75-ohm line sufficiently long to provide for rotation of the antenna, and then can be converted to a balanced 300-ohm load at the anchor point by using a balun as shown in Fig. 10-10.

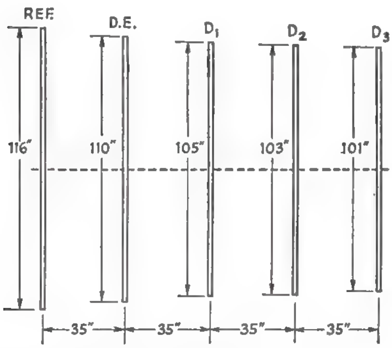


Fig. 10-18—Five-element Yagi to fit on a 12-foot boom. Dimensions are centered at 51 Mc. for working over the 50-52-Mc. range.

Elements should be spaced approximately 0.15 wavelength, or about 35 inches. With 5 or more elements, good band width can be secured by tapering the element lengths properly. With the dimensions given in Fig. 10-18 the antenna will work well over the first two megacycles of the band, provided that the s.w.r. is adjusted for minimum at 51 Mc.

A Dual Array for 28 and 50 Mc.

As many 50-Mc. enthusiasts also operate on 28 Mc., it is often desirable to stack arrays for the two bands on a common tower and rotating device. Such a dual array, combining a 4-element system for 50 Mc. with a 3-element array for 28 Mc., is shown in Fig. 10-19.

If space limitations make it absolutely necessary, the two arrays may be mounted with but a few inches separating them, but experience has shown that some effectiveness is sacrificed,

Fig. 10-19—An example of stacking two arrays for different bands on a common support. All-metal construction is employed in this dual array for 10 and 6 meters.

particularly in the array for the higher frequency. A separation of at least three feet is recommended as the minimum for avoiding harmful interaction. In the example shown the separation is six feet, at which distance each antenna performs equally as well as it would if mounted alone.

In this dual array all-metal construction is employed, doing away with the use of insulators in mounting the elements. The booms are made of two pieces of 1-inch angle stock (24ST aluminum), with supporting braces of the same material. The method of assembling the booms and mounting the elements is shown in Fig. 10-20. The booms are 150 inches and 160 inches in length for the 6- and 10-meter arrays respectively. To prevent swaying of the 10-meter elements, they are braced with guy wires, which are broken up with small insulators. These sway-brace wires are attached to the elements at approximately the midpoint between the boom and the outer end, and are brought up to the vertical support at the point of attachment of the horizontal fore-and-aft braces.

The 50-Mc. antenna is similar in element length and spacing to the 4-element array of Fig. 10-17. The element spacing for the 10-

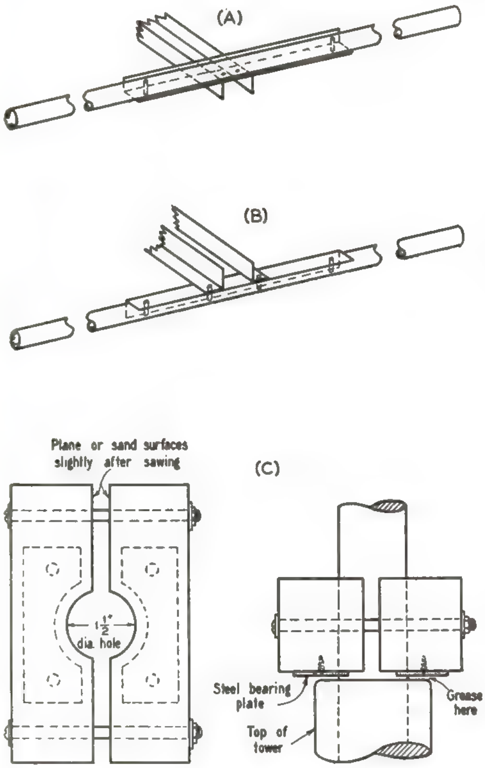


Fig. 10-20—Detail sketches of portions of the dual array for 10 and 6 meters. A — The 50-Mc. boom is made of two pieces of angle stock mounted edge to edge to form a channel. Elements are cradled in another piece of angle stock. B — The two sides of the 10-meter boom are separated and mount on either side of the vertical support. The elements and their supporting crossarms are attached to the lower surface of the boom. C — The bearing for the array is made from a block of wood, drilled to the pipe size, and then sawed lengthwise. It is faced with two steel plates where it rests on the top of the tower.

meter array is 0.2 wavelength, or 80 inches. The driven element is 198 inches long, the director 188 inches, and the reflector 208 inches. It is fed by means of a T match and a 300-ohm line. Clips are set at about 25 inches each side of center. These dimensions give quite uniform performance and low standing-wave ratio over the range from 28 to 29.1 Mc.

WIDE SPACED 6-ELEMENT YAGI

High gain can be obtained by extending the length of a Yagi array to several wavelengths, as discussed in Chapter Four. On 50 Mc. an overall length of about two wavelengths, requiring a boom of the order of 20 feet long, is practicable constructionally. Fig. 10-21 shows a 6-element array of this type, and Fig. 10-22 gives the element dimensions and spacings for a center frequency of 50.6 Mc. The same element arrangement can be used for frequencies higher in the band by subtracting 2 inches from every dimension for each megacycle increase in center frequency. The band width for 2 to 1 s.w.r. is approximately 1 megacycle.

The elements in the antenna shown in Fig. 10-21 are half-inch aluminum tubing, mounted on a 1¼-inch diameter dural boom by the use of HASL castings of the type used in the 3-element array described earlier (Fig. 10-13). If a boom of the requisite length cannot be obtained in one piece it can be made by splicing two or more pieces together, if suspension bracing such as is shown in Fig. 10-21 is used. The boom in this case was made from three pieces of tubing with a short length of the next smaller diameter tubing inside the joint. As the fit was loose, the joint was shimmed with flat strips of sheet aluminum. After assembly a few sheet-metal screws were used to make a tight joint. The element-mounting castings, if used, should be installed before these screws are put in place.

Joints in the boom cause no serious problem if the suspension shown is used to provide additional support. Steel wire can be substituted for the tubing that was incorporated in this antenna. Fig. 10-23 shows the method of fastening the tubing to the pipe mast. A comparable clamp can be used at the boom end of the brace.

Fig. 10-23 also shows an alternative method of mounting the elements on the boom. The clips

Fig. 10-21—Six-element long Yagi for 50 Mc. Boom has two braces running diagonally to the mast, a method of construction that permits using tubing of relatively small diameter and obviates the necessity for a one-piece boom. (The antenna at the top is a 16-element broadside array for 144 Mc.)

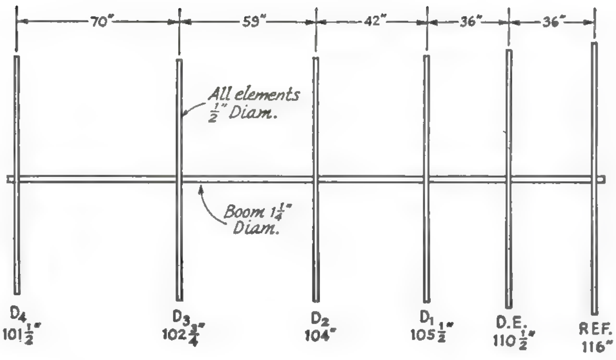


Fig. 10-22—Dimension drawing of the 6-element 50-Mc. antenna. The driven element is fed through a gamma matching section constructed as shown in Fig. 10-24.

should be formed so that when the bolt is tightened the element is pulled tight against the bottom edges of the holes in the boom.

The boom of this antenna is fastened to the pipe mast by the method shown in Fig. 10-16, except that the plate was made long enough so that four U clamps could be used on the boom.

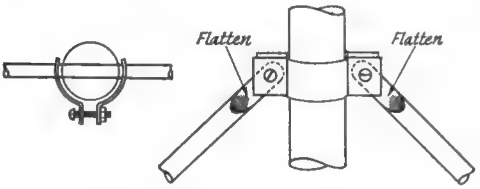


Fig. 10-23—Method of mounting suspension braces to a pipe mast is shown at the right. The braces are flattened at the ends and bolted to semicircular clamps formed from aluminum sheet. A method of mounting elements on a boom is shown at the left. Elements go through holdes drilled through the boom, and are held tight by the pull-down clamps as shown. The clamps can be made from sheet aluminum.

The 6-element antenna is fed with 52-ohm coax through the gamma match shown in Fig. 10-24. This uses a tubular capacitor of the type described in Chapter Twelve. The outside tube is the same material as that used for the elements, and is supported from the driven element by ceramic stand-off insulators one inch high. These are fastened to the tubing with clips made from sheet aluminum. The sliding arm is ¼-inch rod insulated from the outer tube by polystyrene bushings. One of these is force-fitted on the sliding rod with its outer diameter such as to make a sliding fit inside the tube. The other bushing fits tightly inside the outer end of the tube and is drilled at the center so the sliding rod can move freely through it. The bushings can be made from poly rod.

The matching section should be adjusted by using a standing-wave ratio bridge, with the bridge preferably at the antenna during the adjustment process. The antenna can be temporarily mounted eight or ten feet above ground while this adjustment is being made if the feed point is not accessible in the final location.

144-MC. PARASITIC ARRAYS

The main features of the 50-Mc. arrays previously described can be adapted to 144-Mc. antennas, but the small physical size of arrays for this frequency makes it possible to use larger numbers of elements with ease. Few 2-meter antennas have less than 4 or 5 elements, and most stations use more, either in a single bay or in stacked systems.

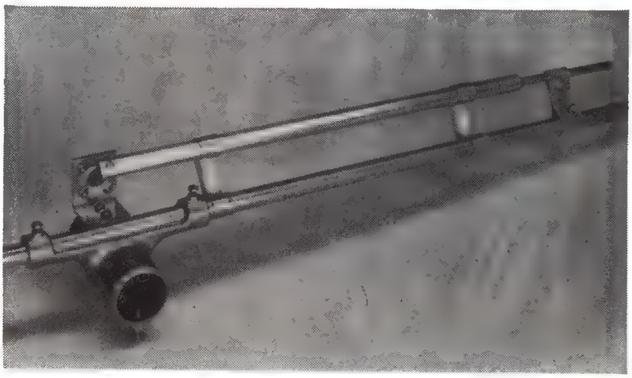


Fig. 10-24—Gamma matching section using tubular capacitor. The sheet-aluminum clip at the right is moved along the driven element for matching. The small rod can be slid in and out of the 15-inch tube for adjustment of series capacitance. The rod should be abcut 14 inches long.

A 4-Element Array

Parasitic arrays for 144 Mc. can be made readily from TV antennas for Channels 4, 5, or 6. The relatively close spacing normally used in TV arrays makes it possible to approximate the recommended 0.2 wavelength at 144 Mc., though the element spacing is not a critical factor. A 4-element array for 144 Mc. that can be made from a Channel 6 TV Yagi is shown in Fig. 10-25. It may be fed with a gamma match and 52-ohm coax, as shown. However, most TV antennas are designed for 300-ohm feed, and the same feed system can be employed for the 2-meter array that is made from them.

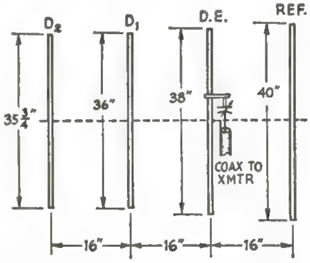


Fig. 10-25—Four-element 144-Mc. Yagi. Gamma matching is recommended, using a gamma section of similar construction to that shown in Fig. 10-15, with a gamma rod having a length of 6 inches. The series capacitor should be a 50-pf. variable; receiver-type plate spacing is adequate for power levels up to a few hundred watts.

If one wishes to build his own Yagi antennas from available tubing sizes, the boom of a 2-meter antenna should be ¾ to 1 inch aluminum or dural. Elements can be ¼ to ½-inch stock. They can be fastened to the boom by the method shown in Fig. 10-23 if the relative diameters of elements and boom are such that the boom will not be unduly weakened by this construction. An alternative mounting method is shown in Fig. 10-30.

Recommended spacing for up to 6 elements is 0.2 wavelength, though this is not too critical. Gamma match feed is recommended for coax, or a folded dipole and balun may be used. If balanced line is to be used the folded dipole is recommended, the 4 to 1 ratio of conductor sizes being about right for most designs.

LONG YAGIS FOR 144 MC.

It becomes practicable, constructionally, at 144 Mc. and higher frequencies to build Yagi antennas that are several wavelengths long, resulting in increased gain and directivity as described in Chapter Four. A representative design is shown in Fig. 10-26. It uses 13 elements —reflector, driven element, and eleven directors—with the director lengths following the third (D_3) successively ⅛ inch shorter. The lengths given apply when the type of element mounting shown in Fig. 10-27 is used—i.e., the element is not mounted through the boom but is fastened to it on top—and when the elements are 3/32 inch in diameter. Steel rods were used as elements in the original model of this antenna. According to the designers, W2NLY and W6QKI, the element diameter should not exceed ⅛ inch.

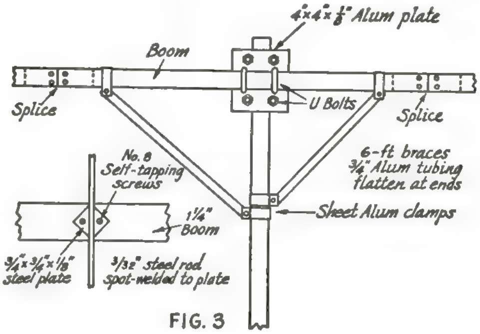
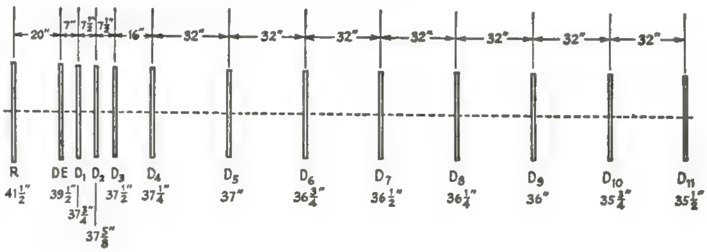


Fig. 10-27—Boom support and element mounting used in the 13-element Yagi antenna of Fig. 10-26.

The driven element (which need not be of the same construction as the reflector and directors) can be a folded dipole or can be fed by any of the matching systems discussed earlier. The feed-point impedance with a simple dipole fed element is 15 to 20 ohms, about the same as in a 3-element Yagi of ordinary design.

Fig. 10-26—Thirteen-element long Yagi antenna for 144 Mc. (W2NLY-W6QKI). Dimensions are for optimum performance in the 144-145-Mc. segment of the band, for maximum performance in higher portions, decrease the element lengths ¼ inch for each megacycle increase in frequency. Dimensions shown apply only for the type of parasitic element construction described in the text and shown in Fig. 10-27.



The length of the antenna can be extended, with some increase in gain (see Chapter Four) by adding similarly-spaced directors, similarly tapered in length. It may also be made shorter, with reduced gain, by cutting off any desired section to the right of D_5 in Fig. 10-26.

In terms of wavelength, the optimum element spacings for an antenna of this type, as determined by W2NLY-W6QKI, are slightly over 0.1 wavelength from driven element to first director and between the first, second and third directors, 0.2 wavelength from third to fourth director, and 0.4 wavelength between succeeding directors.

STACKED YAGI ARRAYS

As discussed in Chapter Four, the gain of a Yagi array can be increased by stacking two (or more) unit arrays or bays and feeding them in phase. Horizontal systems are usually stacked vertically and vertically polarized bays are usually stacked side by side.

Optimum spacing between bays is ¾ wavelength or more, as explained in Chapter Four. Bays of 5 elements or more, spaced one wavelength, are commonly used in antennas for 144 Mc. and higher frequencies. However, at 50 Mc. a spacing of more than ⅝ wavelength is difficult mechanically.

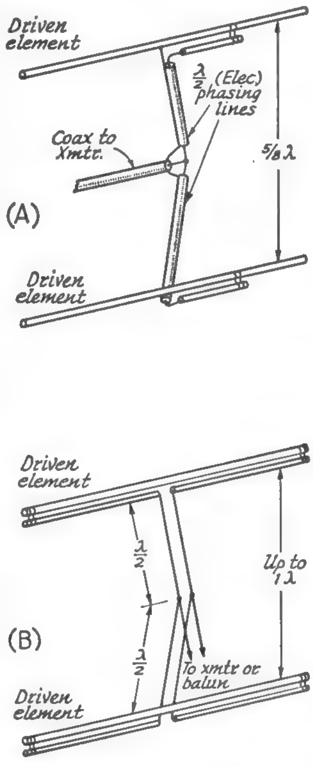


Fig. 10-28 — Two methods for driving parasitic arrays in phase. The method at A can be used for spacings up to nearly ⅔ wavelength with gamma-fed driven elements and coax phasing lines. This length is the limit set by the relationship between electrical length and physical length of solid-dielectric coaxial line. At B, two folded-dipole (or T-matched) driven elements are fed through half-wave sections of open-wire line. Note that in both A and B the lines are connected to the driven elements in exactly the same way: i.e., there is no transposition.

Half-wave stacking is often used in 50-Mc. arrays for convenience in feeding as well as for constructional reasons. Where half-wave stacking is to be employed, the phasing line between bays can be treated as a double Q section. If two bays, each designed for 300-ohm feed, are to be stacked a half wavelength apart and fed at the midpoint between them, the phasing line should have an impedance of about 380 ohms. No. 12 wire spaced one inch will do for this purpose. The midpoint then can be fed either with 300-ohm line or with 72-ohm coax and a balun.

When a spacing of ⅝ wavelength between bays is employed, the phasing lines can be coax, as shown in Fig. 10-28. (The velocity factor of coax makes a full wavelength of line actually about ⅔ wavelength physically.) In this case the total length of coax between bays should be one wavelength, electrically, using the formula given in the section on phasing and matching sections earlier in this chapter for a half wavelength and multiplying by 2. At the midpoint, where the main transmission line connects, the two halves of the phasing line are in parallel, so the impedance at this point is one-half the impedance of each bay. Because of the coupling between the bays, this impedance is somewhat less than half the impedance of the same bay used as a single antenna. The coupling effect is less pronounced with increased spacing between bays.

When two bays are spaced a full wavelength the coupling is relatively slight. The phasing line can be any open-wire line (Fig. 10-28) and the impedance at the midpoint of the line will be approximately half that of the individual bays. Predicting what it will be with a given set of dimensions is difficult, as many factors come into play. It will usually be of a value that can be fed through the combination of a Q section and a transmission line of 300 to 450 ohms impedance. An adjustable Q section like the one shown in Fig. 10-6, may be used when the antenna impedance is not known.

The stacking of Yagis can be carried to the limits of the builder's ability to support and rotate the finished product. Configurations of 6, 8 and even 12 bays have been built successfully for use on 144 Mc.

COLLINEAR ARRAYS FOR 144 MC.

Excellent performance in antenna systems for 144 Mc. and higher bands is obtainable through the use of curtains of 4, 6, 8 or more elements, arranged in pairs and fed in phase. Parasitic reflectors are usually mounted 0.15 to 0.25 wavelength in back of each driven element, though the driven elements alone may be used in a bidirectional array. Screen reflectors are also used with collinear elements. Such arrays may employ either horizontal or vertical polarization. Horizontal is shown in the examples.

The supporting structure may be either wood or metal. If the elements are mounted at their centers (the point of minimum r.f. voltage) no insulation need be used. It is desirable to keep the supporting members in back of the elements, particularly when all-metal construction is employed.

A 12-Element Array

Six half waves in phase, with parasitic reflectors, may be used as shown schematically in Fig. 10-29. The mechanical features of the 12-element array are shown in Figs. 10-30 and 10-31. The spacing of the reflectors in this array is made 0.15 wavelength, to bring down its feed impedance to the point where it can be fed with 300-ohm line without appreciable mismatch. Dimensions are given in the caption for Fig. 10-32.

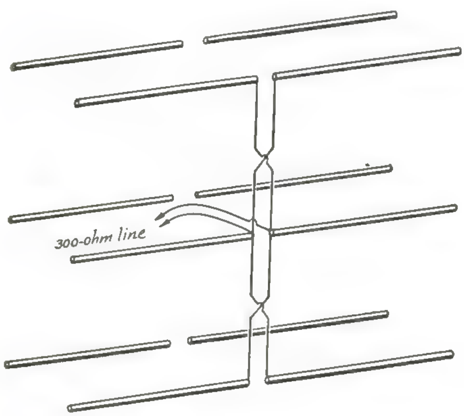


Fig. 10-29—Element arrangement and feed system of the 12-element curtain array. Reflectors are spaced 0.15 wavelength behind the driven elements.

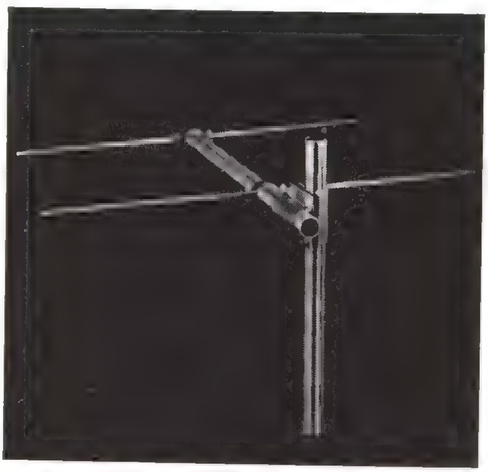


Fig. 10-30—Model showing the method of assembly for all-metal construction of phased arrays. Dimensions of clamps are given in Fig. 10-31.

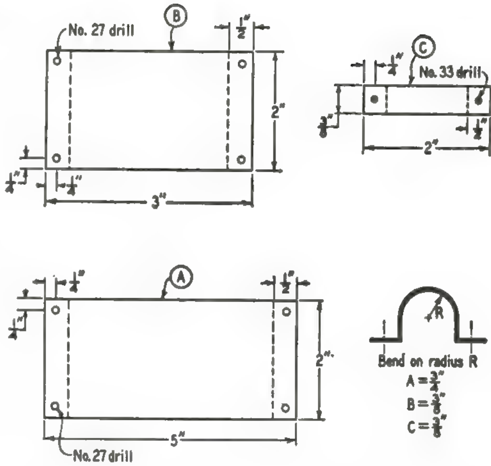


Fig. 10-31—Detail drawings of the clamps used to assemble the all-metal 2-meter array. A, B and C are before bending into "U" shape. The right-angle bends should be made first, along the dotted lines as shown, then the plates may be bent around a piece of pipe of the proper diameter. Sheet stock should be 1/16-inch or heavier aluminum.

A 16-Element Array

Design similar to that given for the 12-element system may be applied to eight half waves in phase, with reflectors, as shown in Fig. 10-33. (This antenna is the uppermost one in Fig. 10-21.) Element dimensions are the same as for the 12-element array, given in the caption for Fig. 10-32.

The extra elements bring the feed impedance of this system down, so the reflector spacing is made 0.2 wavelength for the 16-element array. The impedance is usually slightly below 300 ohms for such an arrangement, though not low enough so that the efficiency is greatly affected. A 300-ohm transmission line may be connected at the midpoint of the phasing line. However, the builder may wish to experiment with an adjustable Q section at the feed point to achieve a more precise match. One feed method that has been employed with the 16-element array is to use a quarter-wave Q section of 300-ohm Twin-Lead, and then make the main transmission line of open-wire line of 400 to 500 ohms impedance.

The Q section may also be any *odd* multiple of a quarter wavelength. This suggests the use of the flexible insulated Twin-Lead as the rotating portion of the transmission line, bringing it to an anchor point just below the array, where open-wire line may comprise the balance of the run. This method is used in the array shown in Fig. 10-21.

Very Large Arrays

Where more than 16 elements are used in a collinear array, the system is usually broken down into separate 12- or 16-element groups,

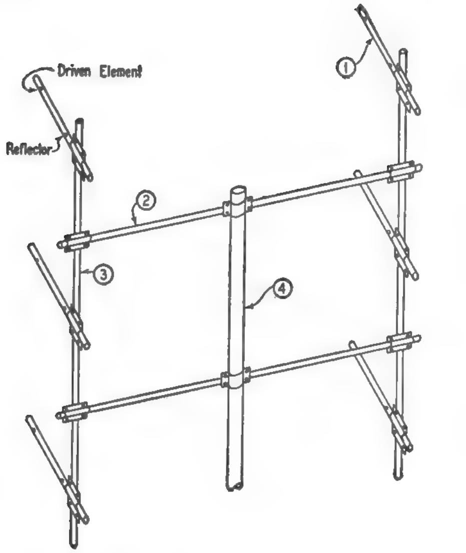


Fig. 10-32—Supporting framework for a 12-element 144-Mc. array of all-metal design. Dimensions are as follows: element supports (1) ¾ by 16 inches; horizontal members (2) ¾ by 46 inches; vertical members (3) ¾ by 86 inches; vertical support (4) 1½-inch diameter, length as required; reflector-to-driven-element spacing 12 inches. Parts not shown in sketch: driven elements ¼ by 38 inches; reflectors ¼ by 40 inches; phasing lines No. 18 spaced 1 inch, 80 inches long, fanned out to 3½ inches at driven elements (transpose each half-wave section). The elements and phasing lines are arranged as shown in Fig. 10-29.

and these are fed in phase. This is done to achieve balanced current distribution. A 24-element array made of two 12-element sets is shown schematically in Fig. 10-34 (reflectors are omitted from the drawing).

A 32-element array built by W1VLH is shown in Fig. 10-35. This comprises two 16-element systems like that in Fig. 10-33. Both it and the 24-element array use one-wavelength phasing lines from each portion, brought to the center of the system and fed in parallel through an adjustable Q section.

The phasing lines can be made any multiple of a half wavelength long, each side of the feed point. This allows the spacing between the sections of the array to be increased, if desired. In the case of the W1VLH array, spacing between the two was adjustable, up to a maximum of ⅝ wavelength between element ends. Wide spacing between the two halves of the array gives somewhat more gain, but is seldom csed because of the greater difficulty of supporting the array.

Element dimensions can be the same in these arrays as in the 12-element antenna of Fig. 10-32. If the array is to be designed for a specific center frequency the charts of Fig. 10-1 should be used for the driven elements. The reflectors should be 5% longer than the driven elements.

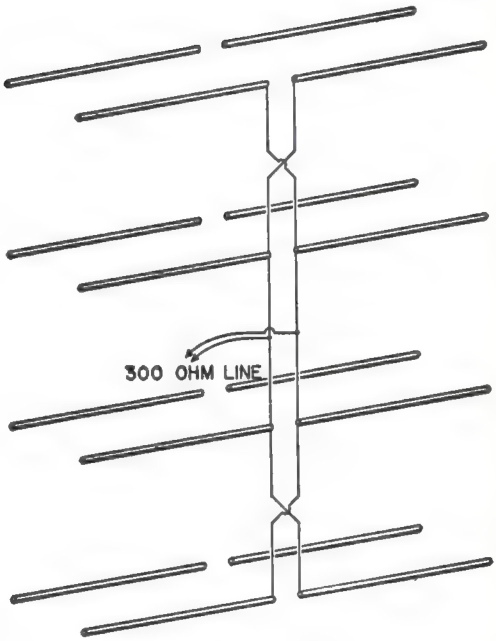


Fig. 10-33—Schematic drawing of a 16-element array. A variable Q section may be inserted at the feed point if accurate matching is desired. Reflector spacing is 0.2 wavelength.

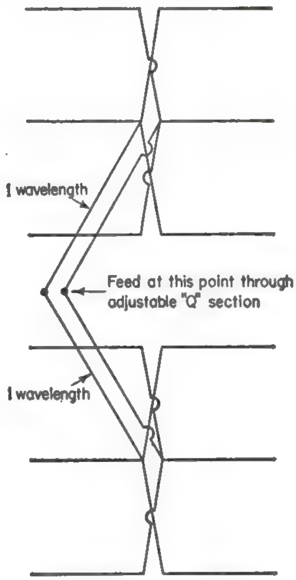


Fig. 10-34—Method of feeding the driven elements of a 24-element array. Phasing lines may be open-wire line with ½ to 1 inch spacing between conductors.

All-Metal Construction

Collinear arrays may be made very light in weight and low in wind resistance, and still have strength to withstand the most severe weather

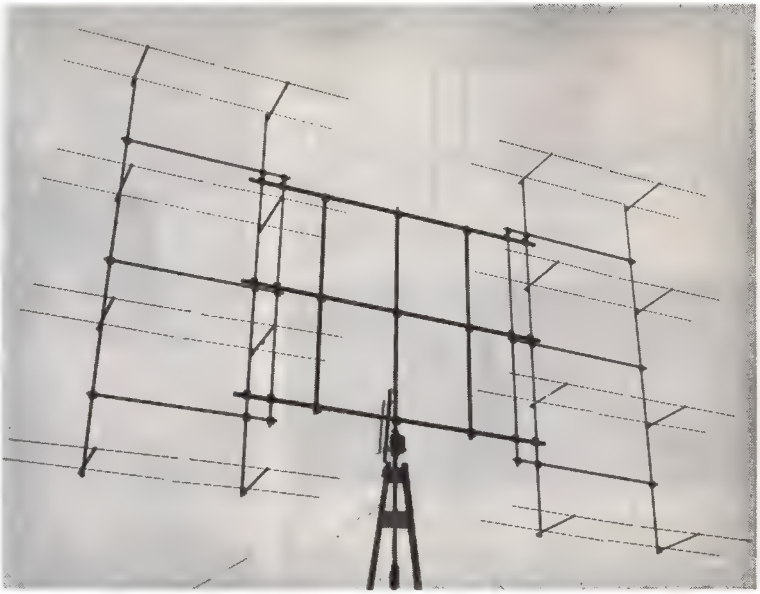


Fig. 10-35—Two 16-element arrays with up to ⅝-wavelength spacing.

conditions, if all-metal designs is employed in the manner shown in Figs. 10-30, 10-31 and 10-32. Elements, supporting arms, and vertical and horizontal supports are all of aluminum or dural tubing, and are held together by clamps made from sheet aluminum. Dimensions for the clamps required when the members are ¼-inch, ¾-inch and 1½-inch tubing are given in Fig. 10-31. A model showing the method of assembling is shown in Fig. 10-30, and the method of assembling a 12-element array is given in Fig. 10-32.

The clamp method of assembly results in a strong structure that will hold its alignment indefinitely. Clamps for combinations of tubing sizes other than those given may be made by bending up clamps experimentally out of soft metal or cardboard, and then using these as templates for cutting and drilling the sheet aluminum. The stock should be 1/16-inch or heavier, and the clamps should be assembled with No. 8 or larger screws. Lock washers should be used under all nuts. Clamps and their screws and nuts should be sprayed with clear lacquer when the assembly is completed. Use two coats for maximum protection.

Interlaced Quad for 6 and 2 Meter

Fig. 10-36 shows an interlaced quad array with 3 elements on 6 meters and 4 elements on 2 meters. The 6-meter elements have the conventional quarter-wavelength sides, while the 2-meter elements are one half wavelength each side, following the so-called X-quad design, as indicated in Fig. 10-37.

The boom is a 6-ft. length of 1¼-inch square aluminum tubing (6063TS) with ⅛-inch wall thickness. The spreader assemblies are spaced evenly along the boom at intervals of 15 inches. In Fig. 10-36, from left to right, are the 6-meter reflector, the 2-meter reflector, the two driven elements on the same set of spreaders, the 2-

Fig. 10-36—The interlaced quad for 50 and 144 Mc. (K8WYU) is a v.h.f. version of multiband quads more often used on 28, 21 and 14 Mc. With the array in this position, the forward lobes of the antenna are toward the right. The array is light enough to be handled by a TV rotator.

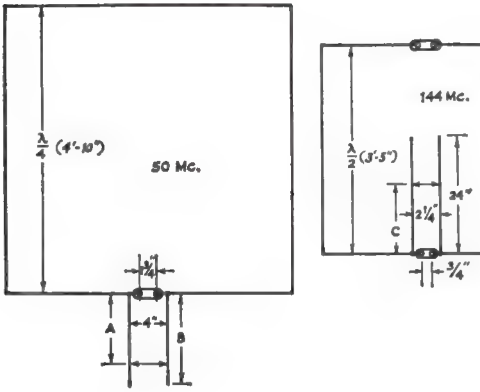


Fig. 10-37—Details of the interlaced quads for 50 and 144 Mc. The tuning stubs on the 6-meter elements hang free, while those on the 2-meter elements are brought up to the center of the array and taped to the boom, as shown in one of the photographs. Approximate dimensions in inches for the various elements are as follows: Director 1—A not used, stub open, B 13 inches, C 11 inches. Director 2—C 12½ inches. Driven element—A and B not used, C 13¼ inches. Reflector—A 18 inches, B 20 inches, C 18¾ inches. All stub dimensions are approximate. Final position of each short is determined with the antenna installed.

meter first director, and the 6-meter director and 2-meter second director both on the same set of spreaders at the right-hand end of the boom.

Each set of spreaders is attached to a square aluminum plate, shown in Fig. 10-38. This plate is fastened to the boom by means of ¼-inch aluminum bolts and four aluminum angles, as shown in Figs. 10-38 and 10-39. Each pair of angles is drilled differently so that the through bolt attaching one pair of angles to the boom will not interfere with the bolt for the second pain.

The sections of ⅞-inch o.d. aluminum tubing, shown in Fig. 10-38 are bolted radially, four to each spreader plate. These serve as socket mountings for spreader arms made of ¾-inch hardwood dowel painted with several coats of Val Oil for weatherproofing. The spreader arms for the 2-meter elements are 29 inches long; those for the 6-meter elements are 41 inches long. The dowel should be inserted the full length of the aluminum tubing to prevent crushing the tubing when the mounting bolts are tightened. Available dowels (at hardware stores or lumber yards) are sometimes limited to lengths of 36 inches. If this is the case, short pieces, 5 inches long, can be inserted in the aluminum tubing sockets first, with the 36-inch length on top of it to make up the 41 inches required for the 6-meter arms.

After the spreaders are assembled, holes can be drilled in the outer ends of the dowels at proper points to give the antenna dimensions shown in Fig. 10-37.

The elements are made of ⅛-inch aluminum clothesline to minimize the weight. Insulator ties at the stubs of the parasitic elements and the feed points of the driven elements are made as shown in Fig. 10-40. A loop is formed in the wire and secured with solder before attaching the wire to the insulator. The aluminum wire can be soldered quite readily by using solder such as Chemalloy (Allied Radio) or Alcoa type 804 with No. 64 flux (both of the latter available from Alcoa distributors). The insulators are made of pieces of polystyrene. After the holes for the wire have been drilled, the piece is sawed lengthwise. The two halves are fitted around the wire of the loops and cemented together with epoxy, and clamped until the cement has hardened. Made in this manner, the insulators are as strong as though they were in one piece, and

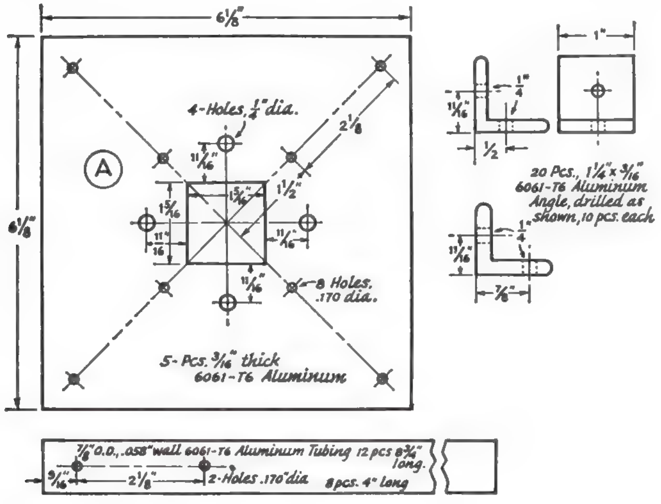


Fig. 10-38—Details of the metalwork for the spiders used to support the elements of the 2-band quad. The square plate fits over the boom, and is held in place by means of four brackets. Pieces of aluminum tubing, below, placed radially on the plate, serve as sockets for the wood-dowel arms which support the wire elements. Those involving the 6-meter elements are 8¾ inches long, while those for the 2-meter elements are 4 inches.

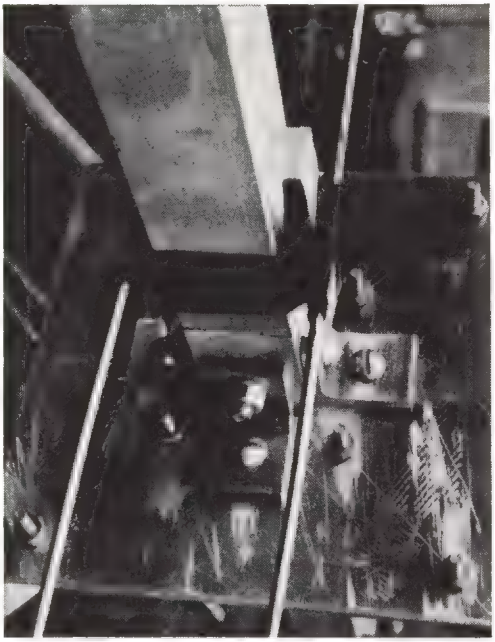


Fig. 10-39—Detail view of one of the 144-Mc. stubs, showing insulating blocks and stub taped in place. One of the square plates for the spider assembly is seen directly in back of the stub.

Fig. 10-40—Close up of one of the element-and-stub assemblies, showing the aluminum soldering. The insulator is made in two pieces and cemented around the wire loops, after soldering is completed and the elements have cooled.

the softening of the insulating material while soldering the wire is avoided.

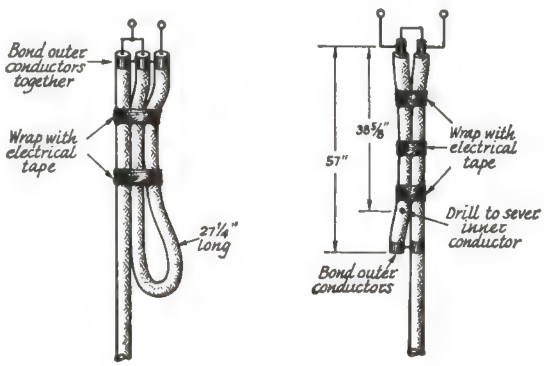
Since copper in contact with aluminum results in destructive chemical action, the transmission lines are terminated in zinc-dipped lugs and these, in turn, should be soldered to the aluminum wire, using ordinary 60/40 solder.

Each of the driven elements is fed with a separate coaxial line of RG-58/CU, although larger cable should be used if the transmitter output power is above 100 watts.

The X quad is voltage fed, so a balun with an impedance step-up is required. This is the conventional half-wave-loop type (left, in Fig. 10-41) commonly used in v.h.f. antenna work. The impedance of the X quad alone would be quite high, but use of large-sized wire and three parasitic elements brings the feed impedance down to the vicinity of 200 ohms. The balun used with this array is made of RG-58/CU cable, cut for 145 Mc. It is wrapped with electrical tape to make it more rigid and durable. The balun loop is 27¼ inches long, If high power is to be used, the balun should be made of larger coax, such as RG-8/U.

The feed impedance of the 50-Mc. antenna is near 50 ohms, so a 1-to-1 balun is required. This is made as shown at the right in Fig. 10-41. A piece of the same 50-ohm cable as is used for the main transmission line is cut to 57 inches over-all, and taped to the line. At the end away from the antenna the outer condcctor of the stub is shorted to the outer conductor of the transmission line with a piece of braid, as shown in the sketch. The inner conductor of the stub is severed at 38⅝ inches from the antenna end, by drilling through the coax. Use care to prevent shorting the inner and outer conductors together in doing this, and be sure that the inner conductor is broken. Check on these points with a continuity meter of some sort, before connecting the balun to the antenna. Note that in its completed form the inner conductors of the stub and the coaxial line are connected together, and the driven-element ends are connected to the outer

Fig. 10-41—The balun used for the 2-meter quad, left, is the conventional half-wave loop, giving a 4-to-1 impedance step-up. The 50-Mc. balun, right, is the 1-to-1 variety. The inner conductor of the stub is severed at the point indicated, by drilling carefully.



conductors. Like the 2-meter balun, this one is wrapped with plastic tape to maintain electrical and mechanical characteristics. The need for larger cable for high-power operation also applies.

The stubs for the parasitic elements are cut to the approximate lengths shown in Fig. 10-37. The stubs of all parasitic elements with the exception of the 6-meter director are closed, adjustment being made by means of a sliding shorting bar. The stub of the 6-meter diector is open, and adjustment is made by altering the length of the stub. The 50-Mc. stubs are allowed to hang free from the elements, but the 144-Mc. stubs are brought up to the boom and taped to blocks of polystyrene or plexiglas, which are taped to the boom, as shown in Fig. 10-39. Once the proper point for the stub shorting bar has been found, the excess stub length may be cut off.

The elements should be adjusted for the center of the frequency range in which most operation will take place. For convenience, a preliminary adjustment may be made in the basement, by suspending the array about a foot below the ceiling beams and about the same distance above the floor.

A grid-dip meter should be used to check the resonant frequency of the parasitic elements. The meter coil is placed alongside the vertical portion of the 50-Mc. element to be adjusted. It is a good idea to check the actual frequency of the g.d.o. by means of a calibrated receiver, as the close coupling needed to get a dip may shift the g.d.o. frequency appreciably. The director is resonated at 54 Mc. by trimming the stub to length, and the reflector to 48 Mc. by adjusting the position of the short.

In tuning the 144-Mc. antenna, the meter coil should be held near the short on the stub, and monitoring of the frequency with a receiver is nearly a necessity. It may be possible to observe two dips, one the correct one and the other resulting from resonance in the open portion of the stub beyond the short. If this occurs, put a temporary short across the open portion of the stub to detune it from the region of the band. A pair of alligator clips connected back-to-back makes a good temporary short. The forward director is tuned to 150 Mc., the second to 149 Mc., the driven element to 145 Mc., and the reflector to 140 Mc.

If final adjustment of the array cannot be made in the position in which it will be used, the next best thing is to put it about 10 feet above ground. The height will be more critical for the 50-Mc. portion than for the higher frequency, and a height above ground of a half wavelength at 50 Mc. will cause the least reaction on the antenna impedance. At this height the stubs can be reached with a stepladder.

Final adjustment is made for maximum front-to-back ratio, using received signals from fairly distant stations. This procedure should be quite straightforward at 50 Mc., but the correct adjustment may not be so easily recognized at 144 Mc. because of reflections from nearby objects. This is less noticeable if the array is elevated 30 ft. or more.

The completed antenna may be sprayed with Krylon as a protection against weather and corrosive gases. The points of connection of the baluns to the antennas should be wrapped carefully with plastic tape, and this wrap may be lacquer sprayed.

LONG-WIRE ARRAYS FOR 50 AND 144 MC.

Long-wire antenna systems such as the V or the rhombic can usually be used on 50 or even 144 Mc. with good results even though they were erected with lower-frequency operation in mind. The included angles in such arrays will not be optimum for v.h.f., but the arrays will be so large, in terms of v.h.f. wavelengths, that they will work well, particularly if the feeder systems are not too long. They will show little frequency discrimination over an entire v.h.f. band.

Long-wire arrays may be constructed according to the principles given in Chapter Five, designing them specifically for v.h.f. use. In such instances an effective rhombic array assumes proportions that make it usable in many amateur locations where a similar array for 7 or 14 Mc. would be out of the question because of its great size. By proper choice of leg lengths a V or rhombic can be made to work well on both 28 and 50 Mc., making it a highly useful system where the necessary space is available for its erection. Several examples are given in Table 10-I.

The tilt angles and leg lengths in wavelengths for other combinations can be worked out from the design data in Chapter Five, using the curves for zero wave angle. The wire lengths in feet are then given by

$$\textit{Length}\ \text{(feet)} = \frac{492\ (N - 0.05)}{\textit{Freq.}\ \text{(Mc.)}}$$

where N is the number of *half waves* on the leg. The above formula need be used only where the leg length is short in terms of wavelength. For longer dimensions the standard half-wave formula may be used:

$$\textit{Length}\ \text{(feet)} = \frac{492N}{\textit{Freq.}\ \text{(Mc.)}}$$

Long-wire systems for combining operation on 50 and 144 Mc. are even more attractive as to size. Because of the nearness to third-harmonic relationship which exists between these two bands, the same matching section and feeder may be used to feed a terminated rhombic for both bands with a flat line. Since a Q section can be any odd multiple of a quarter wavelength, the matching section for a two-band v.h.f. rhombic can be a quarter wavelength long at 50 Mc., in which case it will be approximately three quarter waves long at 144 Mc. The feed impedance of a terminated rhombic is about 800

ohms; thus a 490-ohm Q section is required to match this impedance to a 300-ohm line. Such a matching section could be made of No. 14 wire spaced 1¾ inches, about 53 inches long, as a compromise for the two bands. The array could be fed directly with a 600-ohm line, without appreciable mismatch. Preferably such a line would be of small wire, in order to keep the spacing to relatively small dimensions. See Chapter Three for wire sizes and line impedances.

Laying out a rhombic antenna for the v.h.f. bands is somewhat less complicated than for lower frequencies, because it is usually possible to have the v.h.f. array high enough (in terms of wavelength) so that the effect of ground is a minor consideration. The dimensions given in Table 10-I are based on the assumption that the lowest possible radiation angle is desired, in which case one side should be a half wave longer than half the over-all length. Using the terms of Table 10-I:

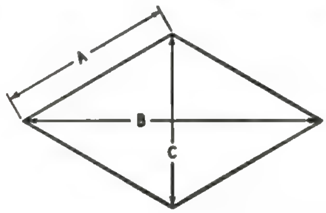
$$A = \frac{B}{2} + \frac{480}{\textit{Freq. (Mc.)}}$$

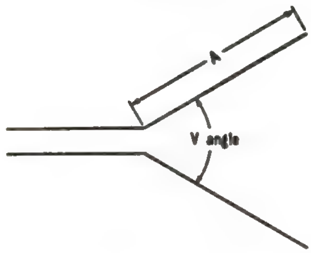
The shape of a multiband V or rhombic may be set up according to Table 10-I with its width, *C*, at the optimum value for the band where highest efficiency is desired. It will be noted that the larger the array the less difference there is in the included angles for adjacent bands. In other words, the larger the array the better will be its capabilities for multiband operation.

TABLE 10-I

Dimensions of V and Rhombic Antennas for V.H.F. Use. Columns 1 and 2 Are for V Designs. For Rhombics Use 1, 3 and 4.

Freq. (Mc.)	*Side Length "A" in Ft.*	*V Angle*	*Over-all Length "B" in Ft.*	*Width "C" in Ft.*
50.5	58	60°	96.5	65.5
145	58	35°	109	39
28.7	68	70°	101.5	84
50.5	68	55°	106.5	70.6
145	68	35°	129	41
50.5	106	42°	192.5	91.5
145	106	35°	205	47.5
28.7	136	52°	237.5	133
50.5	136	37°	252.5	102





Arrays for 220 and 420 Mc.

The use of high-gain antenna systems is virtually a necessity on 220 Mc. and higher frequencies, if communication is to be carried on over other than line-of-sight distances. Experimentation with antenna arrays for these frequencies is fascinating in the extreme, and the size of elements and supporting structures is such that various element arrangements and feed systems can be tried with ease. Arrays for 420 Mc., particularly, are ideal for scale-model demonstrations of antenna principles, as even high gain systems may be of table-top proportions.

Any of the arrays already described may be scaled down for use on 220 and 420 Mc. by reducing all dimensions in proportion to the wavelength, or in inverse proportion to the frequency. Using 144-Mc. designs as a base, the scale factor is 144/220, or 0.655, for converting a 144-Mc. antenna design to 220 Mc., and 144/420, or 0.343, for converting 144 Mc. to 420 Mc. Using the scale factor requires reducing *all* dimensions, including the element diameters. However, if a different length/diameter ratio is used in the elements the proper lengths can be found from Fig. 10-I. In most cases the necessary modification will not be large.

On 220 and 420 Mc. the broad frequency response and ease of adjustment of collinear systems make them more attractive than the more critical Yagi configurations. The use of plane and corner reflectors becomes practical from the standpoint of over-all size, and even parabolic reflectors are usable.

Arrays Back-to-Back

Where both 220 and 420 Mc. are going to be used in one location, it is often possible to mount arrays for the two bands back-to-back, as shown in Fig. 10-42. Here a 16-element array for 220 Mc. and a 24-element for 420 Mc. are mounted on the same upright member. The 220-Mc. portion follows the 16-element design shown earlier for 144 Mc. It is fed at the center with tubular 300-ohm Twin-Lead, matched to the center impedance of the array through a quarter-wave Q section of 7/16-inch tubing spaced about 1½ inches center-to-center. This spacing may be adjusted for lowest s.w.r., permitting the array to be fed with close-spaced open-wire lines of up to 450 ohms impedance.

Elements in the 220-Mc. array are of 7/16-inch aluminum fuel-line tubing, which is very light in weight and easily worked. Supporting members are of ¾-inch tubing, assembled with

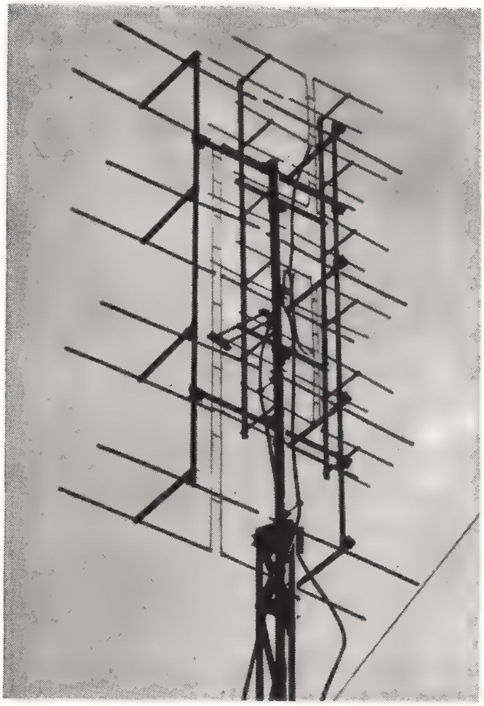


Fig. 10-42—A 24-element array for 420 Mc. and a 16-element for 220 mounted back-to-back on a single support.

clamps similar to those used in the 144-Mc. arrays.

The 420-Mc. array is composed of two 12-element sections, fed as shown in Fig. 10-34. The two portions are joined by phasing sections of 300-ohm Twin-Lead each one wavelength long (21¾ inches). The junction of these presents an impedance of about 150 ohms, and it is fed through an adjustable Q section.

Elements are made of thin-walled ¼-inch tubing, supported in arms of the 7/16-inch tubing used for the 220-Mc. elements. Slots were cut in the ends of these arms to take the elements, and a 4-40 screw was run through both pieces and drawn up tightly with a nut. The arms, in turn, are fastened in holes drilled in the vertical members, and held in place with 6-32 screws and nuts. Because of their small size and light weight, the 420-Mc. elements and supports did not require clamp assembly for strength. Dimensions for both arrays may be taken from Fig. 10-1.

SCREEN-REFLECTOR ARRAYS

At 220 Mc. and higher, where their dimensions become practicable, plane-reflector arrays are widely used. Except as it affects the impedance of the system, as shown by the curve marked 180° in Fig. 10-43, the spacing between the driven elements and the reflecting plane is not particularly critical. Maximum gain occurs around 0.1 to 0.15 wavelength, which is also the region of lowest impedance. Highest impedance appears at about 0.3 wavelength. A plane reflector spaced 0.22 wavelength in back of the driven elements has no effect on their feed impedance. As the gain of a plane-reflector array is nearly constant at spacings from 0.1 to 0.25 wavelength, it may be seen that the spacing may be varied to achieve an impedance match.

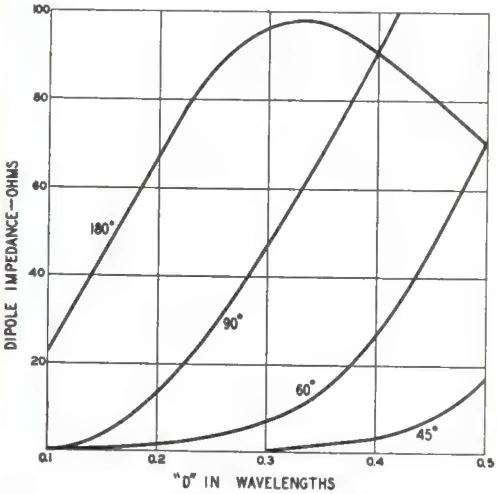


Fig. 10-43—Feed impedance of the driven element in a corner-reflector array, for various corner angles of 180 (flat sheet), 90, 60 and 45 degrees.

An advantage of the plane reflector is that it may be used with two driven-element systems, one on each side of the plane, providing either for two-band operation or the incorporation of horizontal and vertical polarization in a single structure. The gain of a plane-reflector array is slightly higher than that of a similar number of driven elements backed up by parasitic reflectors. The plane-reflector array also has a broader frequency response and higher front-to-back ratio. To achieve these ends, the reflecting plane must be larger than the area of the driven elements, extending at least a quarter wavelength on all sides. Chicken wire on a wood or metal frame makes a good plane reflector. Closely spaced wires or rods may be substituted, for lower wind resistance, with the spacing between them running up to 0.1 wavelength without appreciable loss in effectiveness.

The Corner Reflector

When a single driven element is employed, the plane reflector may be bent to form an angle, giving an improvement in the radiation pattern and gain. The corner reflector, Fig. 10-44, is of a practical size for 220- or 420-Mc. arrays, and a high degree of performance can be attained with a relatively small array. Corner angles of 60 or 90 degrees provide about opti-

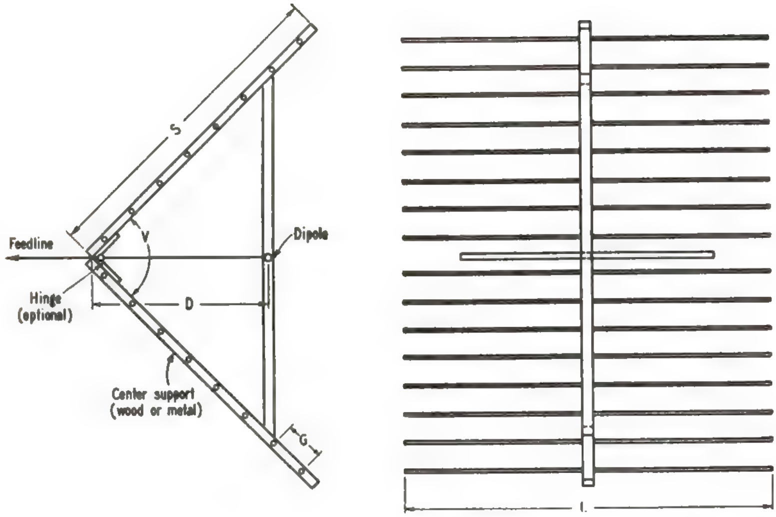


Fig. 10-44—The square-corner array using a grid-type reflector. Elements are stiff wire or tubing. The frame may be either metal or wood. Dimensions are given in Table 10-II.

mum performance, though an angle of 45 degrees may be used, if the side lengths are extended sufficiently. The dipole is parallel to the corner and lies in the plane bisecting the corner angle, as shown in Fig. 10-44. The spacing from the driven element to the corner may be anything from 0.25 to 0.7 wavelength for a 90-degree corner, 0.35 to 0.75 wavelength for a 60-degree corner, and 0.5 to 1.0 wavelength for a 45-degree corner angle.

The center impedance of the driven element in a corner array varies with the corner angle and the position of the radiator with respect to the vertex. From Fig. 10-43 it may be seen that, with a 90-degree corner angle, the center impedance of the driven element approximates the value for a dipole alone at spacings around 0.35 wavelength, rising to about 150 ohms at 0.5 wavelength. The spacing is not critical as to gain. With a 60-degree corner the impedance is about 70 ohms at a spacing of 0.5 wavelength.

Generally speaking, the dimensions of a corner-reflector array are not at all critical, and the frequency characteristics are much better than for a simple parasitic array of anything like the same gain. A gain of about 12db. can be obtained with a 60-degree corner reflector whose sides are about two wavelengths long. Approximately 10 db. gain can be realized from 60- or 90-degree reflectors with a side length of one wavelength. The reflector may be made of sheet metal, wire netting, or a series of spines, there being very little difference in performance provided the spacing of the spines is kept under 0.06 wavelength. This spacing may even be increased to 0.1 wavelength with only a very slight drop in effectiveness. The spines—i.e., the width of the reflector—should be at least 0.6 wavelength long.

A single corner reflector may be used for several bands, provided the spacing of the spines is set up for the highest frequency, and the side and spine lengths for the lowest. Separate dipoles should be used, of course, each being set at the optimum distance from the vertex for the band in question. Separate feed should be used for each dipole.

The corner array is equally effective for either vertical or horizontal polarization. As suggested by W8JK, who developed the corner array for amateur use, the design may be adapted for portable use by hinging the sides at the vertex, as in Fig. 10-44.

For increased gain, the radiator may be two half waves fed in phase, with a corresponding increase in reflector size, or two more more corner-reflector arrays may be stacked vertically or

TABLE 10-II

Dimensions of Corner-Reflector Arrays for 144, 220 and 420 Mc.

Band (Mc.)	*Side Length "S" (Inches)*	*Dipole to Vertex "D" (Inches)*	*Reflector Length "L" (Inches)*	*Reflector Spacing "G" (Inches)*	*Corner Angle "V" (Degrees)*	*Feed Impedance (Ohms)*
144*	65	27.5	48	7¾	90	70
144	80	40	48	4	90	150
220*	42	18	30	5	90	70
220	52	25	30	3	90	150
220	100	25	30	screen	60	70
420	27	8¾	16¼	2⅝	90	70
420	54	13½	16¼	screen	60	70

* Side length and number of reflector elements somewhat below optimum — slight reduction in gain.

The large designs for 220 and 420 Mc. have a gain in excess of 12 db. — intermediate sizes (1 wavelength on each side) about 10 db.

horizontally and fed in phase. The dimensions for corner arrays are available in convenient form in Table 10-11.

The corner reflector is probably the most effective means of developing high gain when only a single dipole is used in the driven portion of the array. The only superior arrangement would involve the use of parabolic reflectors of very large dimensions.

Parabolic Reflectors

A plane sheet may be formed into the shape of a parabolic curve and used with a driven radiator situated at its focus, to provide a highly-directive antenna system. If the parabolic reflector is sufficiently large so that the distance to the focal point is a number of wavelengths, optical conditions are approached and the wave across the mouth of the reflector is a plane wave. However, if the reflector is of the same order of dimensions as the operating wavelength, or less, the driven radiator is appreciably coupled to the reflecting sheet and minor lobes occur in the pattern. With an aperture (open end of the parabola) of the order of 10 or 20 wavelengths, sizes that may be practical for microwave work, a beam width of approximately 5 degrees may be achieved.

A reflecting paraboloid must be carefully designed and constructed to obtain ideal performance. The antenna must be at the focal point. The most desirable focal length of the parabola is that which places the radiator along the plane of the mouth; this length is equal to one half the mouth radius. At other focal distances interference fields deform the pattern or cancel a sizable portion of the radiation.

A 3-Band Log-Periodic Antenna

The antenna shown in Fig. 10-45 covers the complete range of 140 to 450 Mc. with a gain of approximately 6.5 db. over a dipole and with constant impedance and pattern characteristics *vs.* frequency. It is simple and inexpensive to construct, and requires only a single feed line.

The first step in constructing the antenna is to modify some stainless-steel TV strap-type single standoff insulators. These will be used to hold the antenna elements to the booms. The sketch in Fig. 10-46 shows how the straps are modified. The small threaded insert in the standoff usually comes spot welded in three places. If the strap is clamped edgewise in a vise and the in-

TABLE 10-III

Parts list for the Three-Band Antenna

2	10-foot lengths of ½-inch rigid aluminum conduit
32	stainless-steel TV strip-type single standoff insulators (Channel Master 9662)
2	12-foot lengths of ¼-inch diameter aluminum rod *
1	43-inch length of ¼-inch diameter aluminum rod *
32	¼–20 aluminum or cadmium-plated nuts

* Place both 12-foot sections of ¼-inch aluminum rod together and cut in accordance with the following list so that two pieces of each length are obtained (dimensions are in inches): 19½, 17½, 15, 13, 11¾, 10¾, 9¾, 8¾, 8, 7¼, 6½, 6, 5¾, 4½. The 43-inch piece is cut to obtain two 16½ and two 4¾-inch pieces, a total of 32 elements.

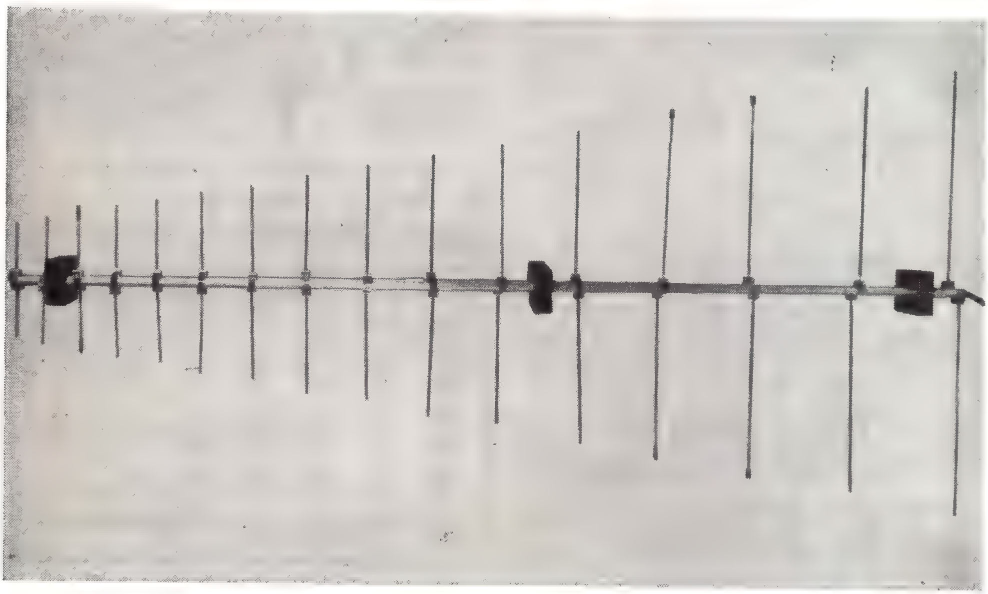


Fig. 10-45—Top view of the three-band log-periodic antenna (K7RTY/2). The three black objects on the booms are the wood block spacers. From this angle only one boom is visible; the other is directly below it.

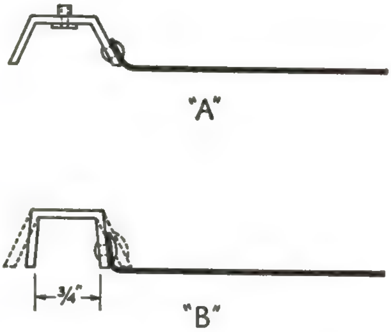


Fig. 10-46—TV strap-type stand-off insulators "A", are modified by removing the threaded insert and by bending the assembly to the dimensions shown in "B."

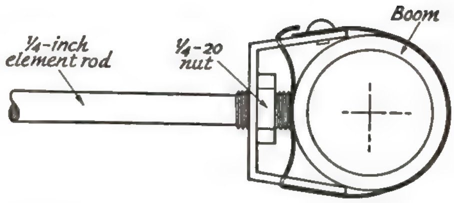


Fig. 10-47—This sketch shows how the elements are attached to the boom.

sert given a sharp rap with a hammer, the insert will fall out without damaging the clamp. Thirty-two straps will be needed for the antenna.

Next, two standard 10-foot lengths of aluminum conduit, ½-inch in diameter, are cut to obtain two 7-foot sections. Aluminum rod, ¼ inch in diameter, is cut to give the required number of elements as shown in Table 10-III. Each rod is then threaded with the ¼-20 die for a distance of about 1 inch on one end.

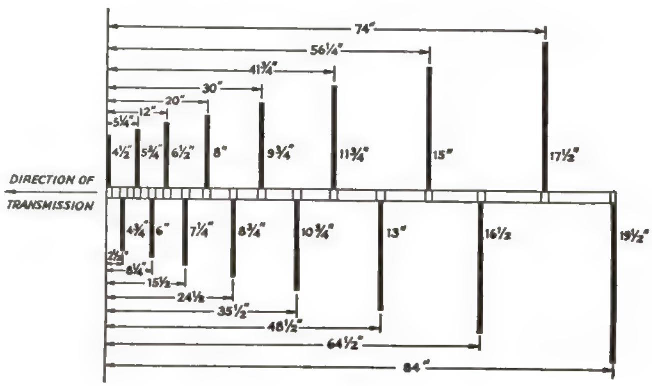
Fig. 10-47 shows the method of attaching elements and clamps to the booms. Fig. 10-48 shows the layout of one section. The complete antenna is made up of two of these sections, one above the other. Any convenient method of clamping the sections together can be used, as long as the booms are insulated from one another. The booms in the antenna shown here were held apart by three wood blocks, shown in Fig. 10-49. Two identical blocks are constructed of 4-inch pieces of 2 ×4 lumber. Two ⅝-inch holes are bored through the blocks and then the blocks are sawed down through the center as shown in Fig. 10-49. The 1¼-inch spacing (between centers) between the two booms should be adhered to as closely as possible. These two blocks are placed near the ends of the antenna booms and clamped together with long bolts or wood screws. The third block is identical to the first two, except that 4 ×4 lumber is used. This permits the use of a mast which can be attached to a suitable coupling mounted on the 4 ×4 center block to support the antenna. All of the blocks should receive several coats of varnish to prevent warping and water absorption.

In assembling the antenna the second section is rotated 180 degrees about the boom axis before it is attached above the first section so that, looking at the completed antenna from the top, the elements of the same length will appear to be end-to-end.

Fig. 10-50 shows the method of feeding the antenna. The coax cable runs through the inside of one boom and is attached to the antenna at the short-element end, as shown. The shield of the coax is folded back and tightened under the clamp which holds the first short element. The center conductor is then run over to the other boom and it, too, is tightened under an element clamp. This method of feeding provides an "infinite balun" and presents a good match to either a 50- or 75-ohm coaxial line.

Radiation patterns measured with the antenna on top of a 50-foot tower show a pattern similar to that of a Yagi with the main lobe off the short element end of the antenna.

Fig. 10-48—Dimensions for one section of the log-periodic antenna. Dimensions along the boom are between element centers. The finished antenna consists of two of these sections, mounted one above the other as shown in the photograph.



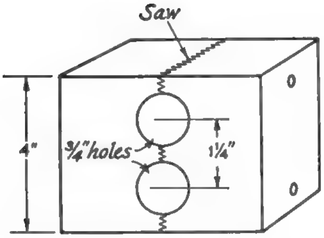


Fig. 10-49—The wood spacer-blocks maintain the proper spacing between the booms. Three blocks are required.

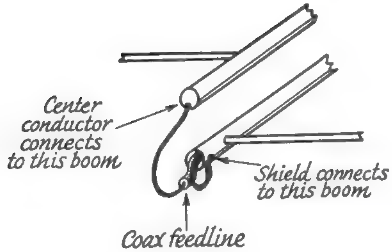


Fig. 10-50—The three-band antenna is fed at the short-element end of the boom. The coax shield connects to one boom and the center conductor connects to the other. The center conductor should be made as short as possible. It is shown here longer than necessary, in to clarify the connections.

Nondirectional V.H.F. Antennas

Most amateur v.h.f. communication is carried on with directive arrays of one kind or another, but in some types of work it is desirable to radiate power equally in all directions. For such work vertical polarization is generally used. Any of the dipole arrangements mentioned earlier in this chapter will give essentially uniform radiation patterns when mounted in a vertical position, but there are modifications that are better adapted to such service.

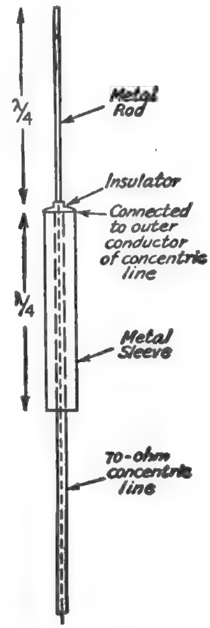


Fig. 10-51—The coaxial antenna. The center conductor of the 70-ohm coaxial line is connected to the rod which forms the upper portion of the antenna, and the outer conductor to the sleeve.

The Coaxial Antenna

End-fed vertical radiators such as the "J" are relatively ineffective because of the tendency of the transmission line to radiate. This condition becomes worse as the frequency increases, becoming quite bad in the v.h.f. range. Radiation from the line combines with that from the antenna itself to raise the radiation angle. To reduce this difficulty the coaxial antenna (Fig. 10-51) was developed. It is used in applications where a non-directional vertical radiator is required.

The center conductor of a 70-ohm concentric transmission line is extended one-quarter wave beyond the end of the line, to act as the upper half of a half-wave antenna. The lower half is provided by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concentric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is nonresonant, since its chracteristic impedance is the same as the center impedance of the half-wave antenna.

Each section of the antenna should be an electrical quarter wave long. The upper rod can be ⅛-inch brass or aluminum and the sleeve can be 1-inch brass or aluminum tubing, in which case suggested dimensions for 144-Mc. work are 19½ inches for the rod and 19¼ inches for the tubing. The insulator can be a few inches of 1-inch diameter polystyrene rod filed down to make a tight fit in the inside of the sleeve, and drilled and tapped at the center for the rod.

A coaxial antenna can be supported from a wooden pole by mounting the sleeve on insulators fastened to the pole, the coaxial transmission line being allowed to drop straight down for at least a half wavelength after leaving the bottom of the sleeve. Alternatively, it can be mounted on top of a pipe mast provided the mast is insulated from the sleeve except where the outside of the coax line connects to it at the top. Thus a supporting pipe of smaller diameter than the sleeve can be run through the center of the sleeve, the coax line being inside the pipe.

Ground Plane Antennas

When a coaxial antena is mounted at the top of a metal mast standing waves may develop on the mast (or on the coaxial cable used to feed the antenna, if the support is nonmetallic) even though the sleeve effects a considerable reduction on the cpupling between the antenna and mast. When this occurs, radiation from the mast combines with that from the antenna to raise the angle of radiation, thereby reducing the effectiveness of the system. The ground-plane type of antenna has largely superseded the coaxial because the horizontal ground plane is an effective shiled between the antenna and mast. The fundamental principles of the ground plane are given in Chapter Two, and its design is discussed in Chapter Three.

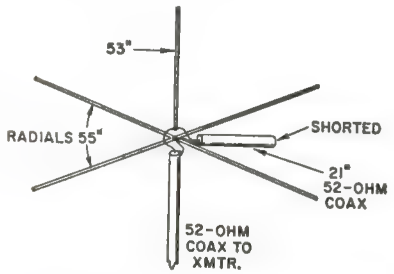


Fig. 10-12—Electrical details of a ground-plane antenna for 50-Mc. work. It will operate with a low s.w.r. over most of the band. If the antenna is supported on a pipe mast (the coax cable can be run through the pipe) a standard floor flange mounted on top of the pipe can serve as a base for supporting the antenna components.

Dimensions for a ground plane antenna designed for the center of the 50-54-Mc. band are given in Fig. 10-52. A matching stub of 52-ohm coax is combined with slight shortening of the vertical radiator to match a 52-ohm feed cable. The dimensions are based on elements having a diameter of ⅛ to ¼ inch, and so are suitable for rod or tubing of readily available sizes. For 144 Mc., using ⅛-inch rods, the dimensions are: radiator (vertical section), 19 inches, radials, 19½ inches each; shorted stub of 52-ohm cable, 1½ inches.

A ground-plane antenna can be constructed from standard plumbing fittings by the method shown in Fig. 10-53. The support is 1-inch (i.d.) from pipe fitted with a standard 1-inch pipe tap. The center of the cap is drilled for a ceramic feed-through bushing as shown, and four equally spaced holes are drilled in the circumference for 1-32 threading. The elements are 19½-inch lengths of ⅛ inch metal rod; either steel or brass can be used (welding or brazing rods usually are easily obtainable). The radiator rod is threaded to go through the insulator with lock nuts on each side, while the radials need be threaded only far enough to seat firmly in the

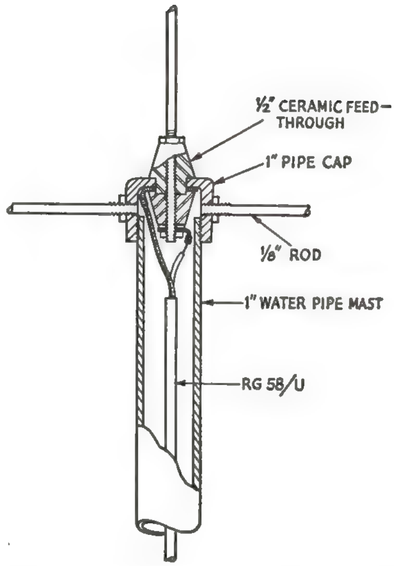


Fig. 10-53—Simple construction for a 144-Mc. ground-plane antenna using standard pipe fitting and readily available components (KØGNP). The coax (52-ohm) shield braid is clamped between the pipe cap and the lower part of the feedthrough insulator, while the center conductor goes to the lower end of the vertical rod.

pipe cap with lock nuts to tighten them in place. In this design no special attempt is made at matching the antenna to the cable, but the standing-wave ratio should not exceed 2 to 1 over a moderate range on either side of the center of the band. An s.w.r. of this order does not materially increase the loss in the cable as compared with perfect matching.

A Collinear Array for 144 Mc.

Where a vertically polarized array having some gain over a dipole is needed, yet directivity is undesirable, collinear halfwave elements may be mounted vertically and fed in phase, as shown in Figs. 10-54 and 10-55. Such an array may have 3 elements, as shown, or 5. The impedance at the center is approximately 300 ohms, permitting it to be fed directly with TV-type line, or through a coaxial balun, as in the model shown. Either 52- or 72-ohm line may be employed without serious mismatch.

The array is made from two pieces of aluminum clothesline wire about 97 inches long overall. These are bent to provide a 38-inch top section, a folded-back 40-inch phasing loop, and a 19-inch center section. These elements are mounted on ceramic pillars, which are fastened to a round wooden pole. Small clamps of sheet aluminum are wrapped around the elements and screwed to the stand-offs. A cheaper but somewhat less desirable method of mounting is to use TV screweye insulators to hold the elements in place.

Feeding the array at the center with a coaxial balun makes a neat arrangement. The balun loop may be taped to the vertical support, and the coaxial line likewise taped at intervals down the mast. The same type of construction can be applied to a 220-Mc. vertical collinear array, using the lengths for that band given in Fig. 10-1.

The vertical support is a round wooden rug or closet pole that can be obtained in any lumber yard. It can be supported in a chimney mount of the type used in TV installations.

Fig. 10-54—Omnidirectional vertical array for 144 Mc. Elements of aluminum clothesline wire are mounted on ceramic standoff insulators screwed to a wooden pole. Feeder shown is 52-ohm coax, with a balun at the feed point. Twin-Lead or other 300-ohm balanced line may also be used, but it should be brought away horizontally from the supporting pole and elements for at least a quarter wavelength. Coax may be taped to the support.

Horizontal Polarization

In regions where horizontal polarization is the rule there is often need for a nondirectional horizontally polarized antenna for local rag-chewing. The "turnstile" type of antenna is capable of radiating a substantially uniform horizontally polarized signal in all azimuthal directions. The electrical arrangement of the turnstile is shown in Fig. 10-56. Two dipoles are crossed at right angles and fed with equal power, but with a phase difference of 90 degrees. The quarter-wave coaxial stub is used to obtain the required phasing of the second element.

Mechanical assembly details of a 144-Mc. turnstile are shown in Fig. 10-57. The support for the elements is a 1-inch length of polystyrene rod 1 inch in diameter. This is drilled as shown in Fig. 10-57 to fit over a ¼-inch rod used as a support but it is readily possible to modify the dimensions for other types of masts. This design uses ⅛-inch dural rods, but brass can be substituted if it is more readily obtainable. Note that none of the rods should be allowed to penetrate the plastic cylinder far enough to make contact with the metal mast. The coax feeder

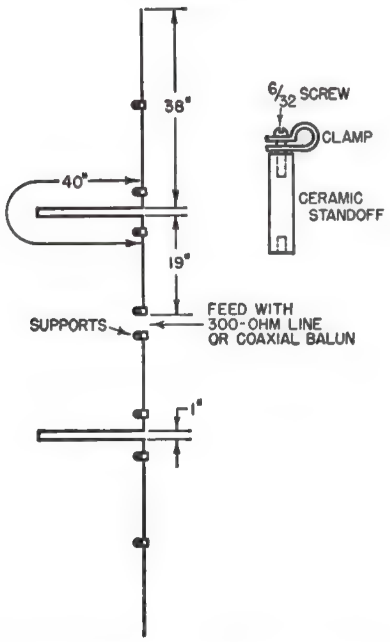


Fig. 10-55—Dimensions and supporting method for the 144-Mc. vertical array.

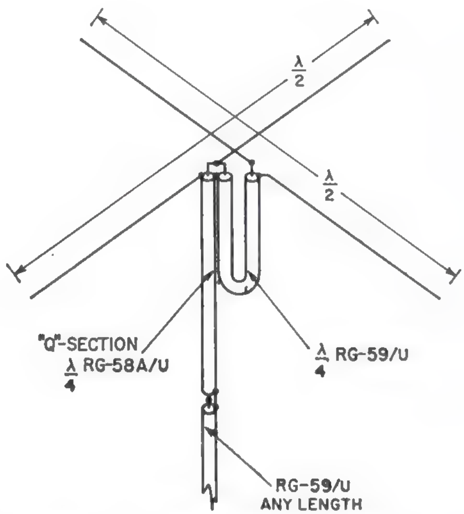


Fig. 10-56—The turnstile antenna. The length of each dipole is calculated by the usual formula: Length in inches = 5540/freq. (Mc.). For 145 Mc. the dipoles are 38¼ inches long. The phasing and Q sections are each 13½ inches for the same frequency.

and phasing section are taped to the support rod in this design, but by using a larger-diameter plastic block and a pipe mast it would be easily possible to run the feed line through the mast.

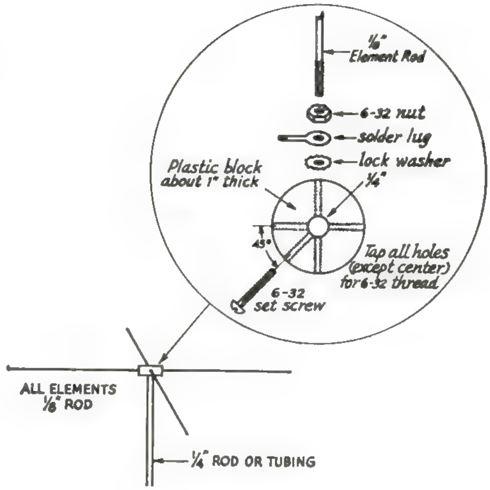


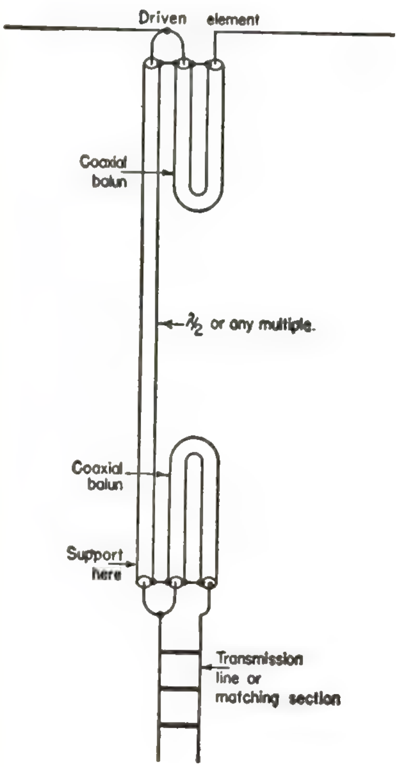
Fig. 10-57—Mechanical details of a light-weight 144-Mc. turnstile antenna.

Practical Construction Hints

Flexible Sections for Rotatable Arrays

When open-wire transmission line is used there is likely to be trouble with shorting or grounding of feeders in rotatable arrays unless some special precuations are taken. Uusually some form of insulated flexible line is connected between the antenna and a stationary support at the top of the tower or mast on which the antenna is mounted.

Such a flexible section can take several forms, and it can be made to double duty. Probably the



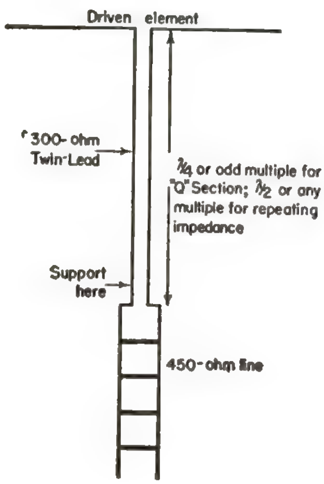


Fig. 10-58—Flexible sections for rotatable arrays. Coax may be used, as at A. If the coax section is any multiple of a half-wave length, the antenna impedance will be repeated at the bottom end. Twin-Lead may be used either as a Q section or as an impedance repeater, as shown in B.

most satisfactory system, for arrays that are not designed to be fed with coaxial line, is to use a flexible section of coax with coaxial baluns at both ends. The outer conductor of the coax may be grounded to the tower or to the beam antenna framework, wherever it is advantageous to do so. Such a flexible section is shown in Fig. 10-58. If the coaxial section is made any multiple of a half wave in electrical length of the impedance of the array will be repeated at the bottom of the flexible section.

Another method is to use Twin-Lead for the flexible section. The 300-ohm tubular type designed for transmitting applications is recommended. Here, again, half-wave sections repeat the antenna feed impedance at the bottom end. The Twin-Lead section may also be made an odd multiple of a quarter wavelength, in which case it will act as a Q section, giving an impedance step-down between a 450-ohm line and an antenna impedance of 200 ohms.

Using TV Antennas in Amateur Work

Ready-made TV antennas can be modified for amateur use with very little trouble. Some antennas designed for TV reception can be used for amateur v.h.f. reception with no change at all.

A Channel 2 Yagi can be used on 50 Mc. as is, though it can probably be improved by lengthening the elements slightly (see Fig. 10-1). The conical TV antenna, being designed for broadband characteristics, will work to some extent on 50, 144 and 200 Mc. Most Channel 6 Yagis can be modified, merely by cutting the element lengths according to Fig. 10-1, to do a good job on 144 Mc. Commercial Yagi arrays for Channels 7 to 13 can be altered for 220-Mc. work by a slight trimming of the elements.

The corner-reflector arrays designed for the u.h.f. TV band will do fairly well on 420 Mc.,

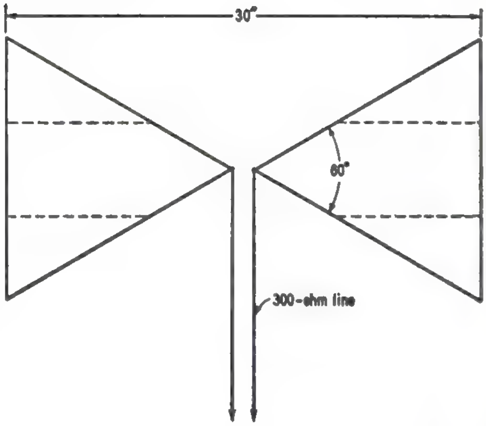


Fig. 10-59—A bow-tie antenna for operation on 220 and 420 Mc. A plane reflector may be added for gain and directivity. Elements may be bent forward on dotted lines for use as a broad-band dipole in a corner reflector.

but their gain will be improved by extending the length of the driven elements slightly. They may be stacked and fed in phase, if the phasing lines are made to 420-Mc. dimensions.

Some ideas can be taken from the TV antenna designer's book for amateur purposes. The u.h.f. "bow-tie" can be made to work on both 220 and 420 Mc. if its triangular dipole elements are made 15 inches long. The angle at the inner points should be 60 degrees. The corner-reflector antenna principle can also be adapted to use on 220 and 420 with a single broad-band dipole, if the dimensions given for the bow-tie are used. The elements should be bent forward, as in the TV corner-reflector systems, approximately as shown in dotted lines in Fig. 10-59.

Chapter 11

Construction of Wire Antennas

Although wire antennas are relatively simple, they can constitute a potential hazard unless properly constructed. Antennas should *never* be run under or over public-utility (telephone or power) lines. Several amateurs have lost their lives by failing to observe this precaution.

Antenna Materials

ANTENNA WIRE

The r.f. resistance of copper wire increases as the size of the wire decreases. However, in most types of antennas that are commonly constructed of wire, the r.f. resistance, even for quite small sizes of wire, will not be so high, compared to the radiation resistance, that the efficiency of the antenna will suffer greatly. Wire sizes as small as No. 30, or even smaller, have been used quite successfully in the construction of "invisible" antennas in areas where there is local objection to the erection of more conventional types. In most cases, the selection of wire for an antenna will be based primarily on the physical properties of the wire, since the suspension of wire from elevated supports places a strain on the wire.[1]

Wire Types

Wire having an enamel-type coating is preferable to bare wire, since the coating resists oxidation and corrosion. Several types of wire having this type of coating are available, depending on the strength needed. "Soft-drawn" or annealed copper wire is easiest to handle but, unfortunately, is subject to considerable stretch under stress. It should therefore be avoided, except for applications where the wire will be under little or no tension, or where some change in length can be tolerated. (For instance, the length of a horizontal antenna fed at the center with open-wire line is not critical, although a change in length may require some readjustment of coupling to the transmitter.)

"Hard-drawn" copper wire or copper-clad steel, especially the latter, is harder to handle, because it has a tendency to spiral when it is unrolled. However, these types are mandatory for applications where significant stretch cannot be tolerated. Care should be exercised in using this wire to make sure that kinks do not develop that may cause the wire to break at far under normal stress. After the coil has been unwound, it is advisable to suspend the wire a few feet above the ground for a day or two before making use of it. The wire should not be recoiled before installing.

The size of the wire to be selected, and the choice between hard-drawn and copper-clad will depend on the length of the unsupported span, the amount of sag that can be tolerated, the stability of the supports under wind pressure, and whether or not an unsupported transmission line is to be suspended from the span.

Wire Tension

Table 11-I shows the maximum rated working tensions of hard-drawn and copper-clad steel wire of various sizes.

The tension on a span of wire can be estimated if one end of the span is attached firmly to one support, while the other end is attached to a halyard running through a pulley fastened to the second support. The tension may then be determined by weighing the load attached to the halyard, and adding the weight of the halyard, if it is significant compared to the weight of the wire. A large bucket loaded with rocks or sand makes a good weight for the purpose.

Wire Splicing

Wire antennas should preferably be made with unbroken lengths of wire. In instances where this is not feasible, wire sections should be spliced as shown in Fig. 11-1. The enamel insulation should be removed for a distance of about 6 inches from the end of each section by

TABLE 11-I
Maximum Recommended Tensions (lbs.)

Wire Size	*Hard-drawn Copper*	*Copper-Clad Steel*[1]
4	214	495
6	130	310
8	84	195
10	52	120
12	32	75
14	20	50
16	13	31
18	8	19
20	5	12

Above tensions are 10% of the breaking load. The figures might be increased 50% if the end supports are firm and there is no danger of icing. [1] 40% copper

[1] The National Electric Code of the National Fire Protection Asssociation contains a section on amateur stations in which a number of recommendations are made concerning minimum size of antenna wire and the manner of bringing the transmission line into the station. The code in itself does not have the force of law, but it is frequently made a part of local building regulations, which are enforceable. The provisions of the code may also be written into, or referred to, in fire and liability insurance documents. A copy of this code may be obtained from National Fire Protection Association, 60 Batterymarch St., Boston, Mass. 02110. (Price $1.00).

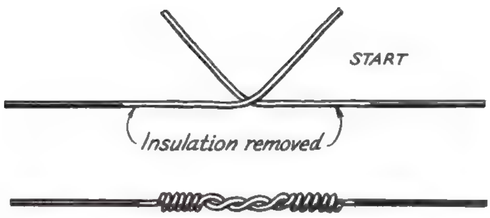


Fig. 11-1—Correct method of splicing antenna wire.

scraping with a knife or rubbing with sandpaper until the copper underneath is bright. The turns of wire should be brought up tight around the standing part of the wire by twisting with broad-nose pliers.

The crevices formed by the wire should be completely filled with solder. Since most antenna soldering must be done outdoors, the ordinary soldering iron or gun may not provide sufficient heat, and the use of a propane torch may become desirable. The joint should be heated sufficiently so that the solder will flow freely into the joint when the source of heat is removed momentarily. After the joint has cooled completely, it should be wiped clean with a cloth, and then sprayed generously with acrylic to discourage corrosion.

ANTENNA INSULATION

To prevent loss of power, the antenna should be well insulated from ground, particularly at the outer end or ends, since these points are always at a comparatively high r.f. potential. If an antenna is to be installed indoors (in an attic, for instance) the antenna may be suspended directly from the wood rafters without additional insulation, if the wood is permanently dry. However, when the antenna is located outside, where it is exposed to wet weather, much greater care should be given to the selection of proper insulators.

Insulator Leakage

The insulators should be of material that will not absorb moisture. Most insulators designed specifically for antenna use are made of glass or glazed porcelain. Aside from this, the length of an insulator in proportion to its surface area is indicative of its comparative insulating abilty. A long thin insulator will have less leakage than a short thick insulator. Some antenna insulators are deeply ribbed to increase the surface leakage path without increasing the physical length of the insulator. Shorter insulators can be used at low-potential points, such as at the center of a dipole. However, if such an antenna is to be fed with open-wire line and used on several bands, the center insulator should be the same as those used at the ends, because high r.f. potential will exist across the center insulator on some bands.

Insulator Stress

As with the antenna wire, the insulator must have sufficient physical strength to sustain the stress without danger of breakage. Long elastic bands, or lengths of nylon fishing line provide long leakage paths and make satisfactory insulators within their limits to resist mechanical strain. They are often used in antennas of the "invisible" type mentioned earlier.

For low-power work with short antennas not subject to appreciable stress, almost any small glass or glazed-porcelain insulator will do. Homemade insulators of Lucite rod or sheet will also be satisfactory. More care is required in the selection of insulators for longer spans and higher transmitter power.

For the same material, the breaking tension of an insulator will be proportional to its cross-sectional area. It should be remembered, however, that the wire hole at the end of the insulator decreases the effective cross-sectional area. For this reason, insulators designed to carry heavy strains are fitted with heavy metal end caps, the eyes being formed in the metal cap, rather than in the insulating material itself. The following stress ratings of several antenna insulators made by E. F. Johnson are typical:

⅝ inch square by 4 inches long—400 lbs.
1 inch diameter by 7 or 12 inches long—800 lbs.
1½ inches diameter by 8, 12 or 20 inches long, with special metal end caps—5000 lbs.

These are rated breaking tensions. The actual working tensions should be limited to not more than 25% of the breaking rating.

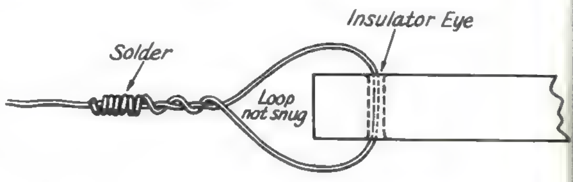


Fig. 11-2—In fastening antenna wire to an insulator, the wire loop should not be made too snug. After completion, solder should be flowed into the turns. When the joint has cooled completely, it should be sprayed with acrylic.

The antenna wire should be attached to the insulators as shown in Fig. 11-2. Care should be taken to avoid sharp angular bends in the wire in looping it through the insulator eye. The loop should be generous enough in size that it will not bind the end of the insulator tightly. If the length of the antenna is critical, the length should be measured to the outward end of the loop, where it passes through the eye of the insulator. The soldering should be done as described earlier for the wire splice.

Strain Insulators

"Strain" insulators have their holes at right angles, since they are designed to be connected

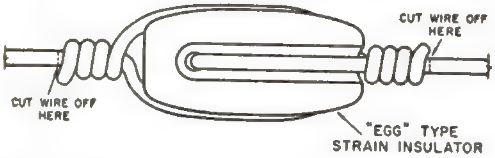


Fig. 11-3—Conventional manner of fastening to a strain insulator. This method decreases the leakage path, and increases capacitance, as discussed in the text.

as shown in Fig. 11-3. It can be seen that this arrangement places the insulating material under compression, rather than tension. An insulator connected this way can withstand much greater stress. Furthermore, if the insulator should break, the wire will not collapse, since the two wire loops are interlocked. However, because the wire is wrapped around the insulator, the leakage path is reduced quite drastically, and the capacitance between the wire loops provides an additional leakage path. For this reason the use of the strain insulator is usually confined to such applications as breaking up resonances in guy wires, where high levels of stress prevail, and where the r.f. insulation is of less importance. Such insulators might, however, be suitable for use at low-potential points on the antenna, such as at the centers of dipoles. These insulators may also be fastened in the conventional manner if the wire will not be under sufficient tension to break the eyes out.

Insulators for Ribbon-Line Antennas

Fig. 11-4A shows the sketch of an insulator designed to be used at the ends of a folded dipole, or a multiple dipole made of ribbon line. It should be made approximately as shown, out of Lucite or bakelite about ¼-inch thick. The advantage of this arrangement is that the strain of the antenna is shared by the conductors and the plastic webbing of the ribbon, which adds considerable strength. After soldering, the screw should be sprayed with acrylic.

Fig. 4B shows a similar arrangement for suspending one dipole from another in a multiple-dipole system.

If better insulation is desired, these insulators can be wired to a conventional insulator.

PULLEYS AND HALYARDS

Pulleys and halyards commonly used to raise and lower the antenna are also items that must be capable of taking the same strain as the antenna wire and insulators. Unfortunately little specific information on the stress ratings of most pulleys is available. Several types of pulleys are readily available at almost any hardware store. Among these are small galvanized pulleys designed for awnings, and several styles and sizes of clothesline pulleys. In judging the stress that any pulley might handle, particular attention should be paid to the diameter of the shaft, how securely the shaft is fitted into the sheath, and the size and material of which the frame is made. Heavier and stronger pulleys are those used in marine work.

Another important factor to be considered in the selection of a pulley is its ability to resist corrosion. Galvanized awning pulleys are probably the most susceptible to corrosion. While the frame or sheath usually stands up well, these pulleys usually fail at the shaft, which eventually rusts out, allowing the grooved wheel to break away under tension.

Most good-quality clothesline pulleys are made of alloys which do not corrode readily. Since they are designed to carry at least 50 ft. of line loaded with wet clothing in stiff winds, they should be adequate for normal spans of 100 to 150 ft. between stable suports. One of the more recent clothesline pulleys has a 4-inch diameter plastic wheel with a ¼-inch shaft in bronze bearings. The sheath is of cast or forged corrosion-proof alloy. Such pulleys sell for about one dollar in hardware stores.

Marine pulleys have good weather-resisting qualities, since they are usually made of bronze, but they are comparatively expensive and are not designed to carry heavy loads. For extremely long spans, the wood-sheathed pulleys used in "block and tackle" devices, and for sail hoisting should fill the requirements.

Halyards

Table 11-II shows recommended maximum tensions for various sizes and types of line and rope suitable for hoisting halyards. Probably the best type for general amateur use for spans up to 150 or 200 ft. is ¼-inch nylon rope. It is somewhat more expensive than ordinary rope of the same size, but it weathers much better. Furthermore, it has a certain amount of elasticity to accommodate gusts of wind, and is particularly

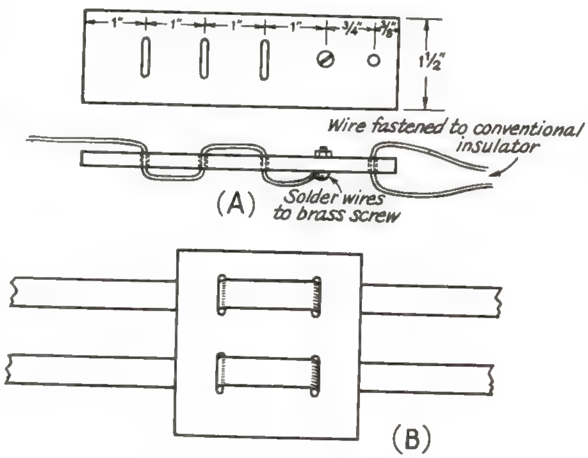


Fig. 11-4—A—Insulator for ends of folded dipoles, or multiple dipoles made of 300-ohm ribbon. B—A method of suspending one ribbon dipole from another in a multiband dipole system.

TABLE 11-II
Approximate Safe Working Tension (lbs.) for Various Halyard Materials

Manila Rope	¼″ – 120 ⅜″ – 270 ½″ – 530 ⅝″ – 800
Polypropylene Rope	¼″ – 270 ⅜″ – 530 ½″ – 840
Nylon Rope	¼″ – 300 ⅜″ – 660 ½″ – 1140
7 x 11 Galvanized Sash Cord	1/16″ – 30 ⅛″ – 125 3/16″ – 250 ¼″ – 450
High-Strength Stranded Galvanized Guy Wire	⅛″ – 400 3/16″ – 700 ¼″ – 1200
Rayon-filled Plastic Clothesline	7/32″ – 60 to 70

recommended for antennas using trees as supports.

Most types of synthetic rope are slippery, and some types of knots ordinarily used for rope will not hold well. Fig. 11-5 shows a knot that should hold well, even with nylon rope or plastic line.

For exceptionally long spans, stranded galvanized steel sash cord is suitable. Cable advertised as "wire rope" usually does not weather well.

A convenience in antenna hoisting (usually a necessity with metal halyards) is the boat winch sold at marinas, and also at such places as Sears.

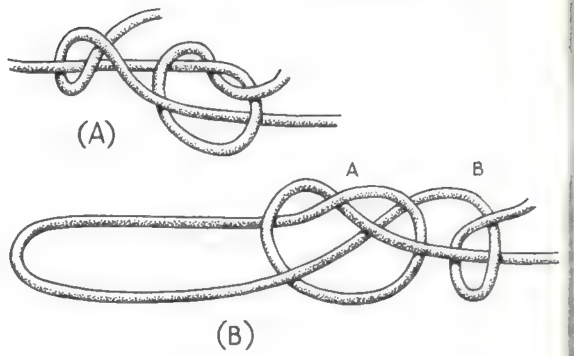


Fig. 11-5—This is one type of knot that will hold with smooth rope, such as nylon. A shows the knot for splicing two ends. B shows the use of a similar knot in forming a loop, as might be needed for attaching an insulator to a halyard. Knot A is first formed loosely 10 or 12 inches from the end of the rope; then the end is passed through the eye of the insulator and knot A. Knot B is then formed and both knots pulled tight. (K7HDB).

Installing Transmission Lines

CONNECTING LINE TO ANTENNA

In connecting coaxial cable or 300-ohm ribbon line to a dipole that does not have a support at the center, it is essential that the conductors of the line be relieved of the weight of the cable or ribbon. Fig. 11-6 shows a method of accomplishing this with coaxial cable. The cable is looped around the center antenna insulator, and clamped before making connections to the antenna. In Fig. 11-7, the weight of the ribbon line is removed from the conductor by threading the line through a sheet of insulating material. The sheet is suspended from the antenna by threading the antenna through the sheet. This arrangement is particularly suited to folded dipoles made of 300-ohm ribbon.

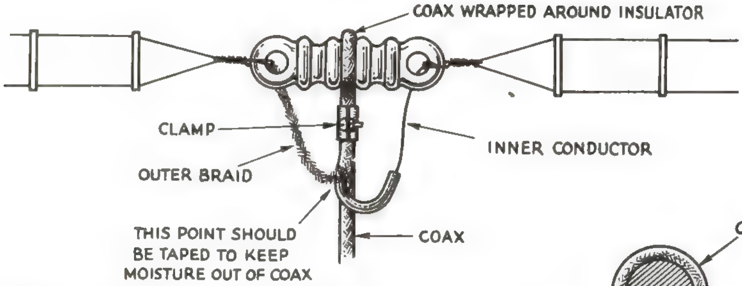
In connecting an open-wire line to an antenna, the conductors of the line should be anchored to the insulator by threading them through the eyes of the insulator two or three times, and twisting the wire back on itself before soldering fast. A slack tie wire should then be used between the feeder conductor and the antenna, as shown in Fig. 11-8. (The tie wires may be extensions of the line conductors themselves.)

When using TV-type open-wire line, the tendency of the line to twist and short out close to the antenna can be counteracted by making the center insulator of the antenna longer than the spacing of the line, as shown in Fig. 11-8. In this case, a heavier spreader insulator should be added just below the antenna insulator to prevent side stress from pulling the conductors away from the light plastic feeder spreaders.

RUNNING LINE FROM ANTENNA TO STATION

Coaxial cable requires no particular care in running from the antenna to the station entrance, except to protect it from mechanical damage. If the antenna is not supported at the center, the line should be fastened to a post more than head high located under the center of the antenna, allowing enough slack between the post and the antenna to take care of any movement of the antenna in the wind. If the antenna feed point is supported by a tower or mast, the cable can be taped at intervals to the mast, or to one leg of the tower.

If desired, coaxial cable can be buried a few inches in the ground in making the run from the antenna to the station. A deep slit can be cut by pushing a square-end spade full depth into the ground, and weaving the handle back and forth to widen the slit before removing the spade. After the cable has been pushed into the slit with a piece of 1-inch board 3 or 4 inches wide, the slit can be closed by tamping.



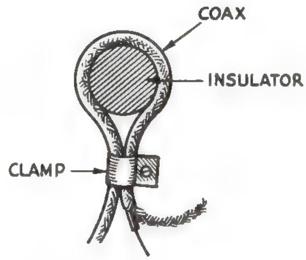


Fig. 11-6—Method of relieving strain on conductors of coaxial cable in feeding a dipole.

Ribbon line should be kept reasonably well spaced from other conductors running parallel to it for more than a few feet. TV-type standoff insulators with strap-clamp mountings can be used in running this type of line down a mast or tower leg. Similar insulators of the screw type can be used in supporting the line on poles for a long run.

Open-wire lines, especially TV types, require frequent supports to keep the line from twisting and shorting out, as well as to relieve the strain. One method of supporting a long run of heavy open-wire line is shown in Fig. 11-9. The line must be securely anchored at a point under the feed point of the antenna. TV-type line can be supported similarly by means of wire links fastened to the insulators. Fig. 11-10 shows a method of supporting an open-wire line from a tower.

To keep the line clear of pedestrians and vehicles, it is usually desirable to anchor the feed line at the eave or rafter line of the station building (see Fig. 11-11), and then drop it vertically to the point of entrance. The points of anchorage and entrance should be chosen so as to permit the vertical drop without crossing windows. If the station is located in a room on the ground floor, one way of bringing the transmission line in is to go through the outside wall below floor level, feed it through the basement, and then up to the station through holes in the floor. In making the entrance hole in the side of the building, suitable measurements should be made in advance to make sure that the hole will go through the sill 2 or 3 inches above the foundation line (and between joists if the bore is parallel to the joists).

The line should be allowed to sag below the entrance-hole level to allow rain water to drip off, and not follow the line into the building.

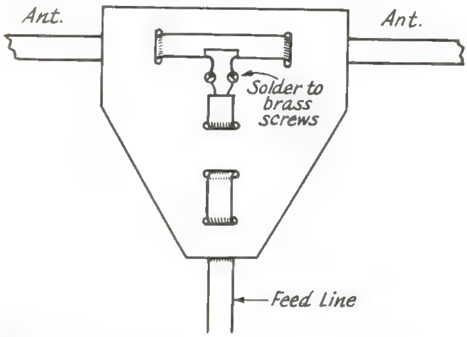


Fig. 11-7—Strain reliever for conductors of 300-ohm ribbon line in a folded dipole. The piece can be made from ¼-inch Lucite sheet.

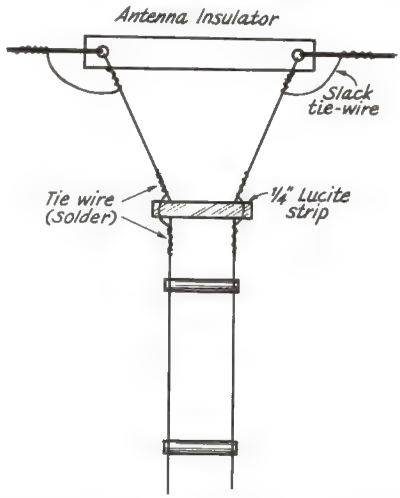


Fig. 11-8—Method of connecting open-wire line to center antenna insulator. The Lucite strip keeps the feedline conductors from pulling away from the spreaders when TV open-wire line is used.

Open-wire line can be fed in a similar manner, although it will require a separate hole for each conductor. The holes should be insulated with lengths of polystyrene or Lucite tubing, and should be drilled with a slight downward slant toward the outside of the building to prevent rain seepage. With TV-type line, it will be necessary to remove a few of the spreader insulators, cut the line before passing through the holes (allowing enough length to reach the inside), and splice the remainder on the inside.

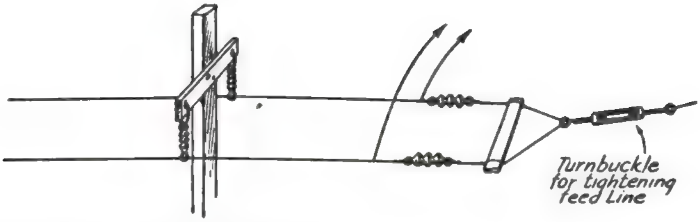


Fig. 11-9—A support for open-wire line. The support at the antenna end of the line must be sufficiently rigid to stand the tension of the line.

If the station is located above ground level, or there is other objection to the procedure described above, entrance can be made at a window, using the arrangement shown in Fig. 11-12. An Amphenol type 83-IF connector can be used as a feedthrough for coaxial line, or one can be made as shown in Fig. 11-13; ceramic feedthrough insulators can be used for open-wire line. Ribbon line can be run through clearance holes in the panel, and secured by a winding of tape on either side of the panel, or by cutting the retaining rings and insulators from a pair of TV standoff insulators, and clamping one on each side of the panel.

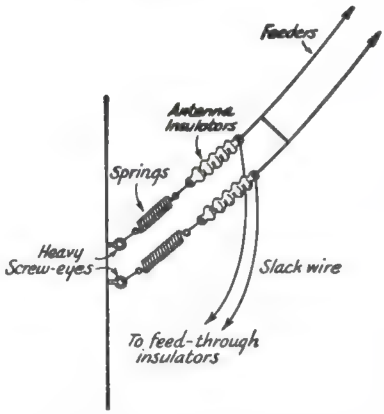


Fig. 11-11—Anchorage for open-wire line at the station end. The springs are especially desirable if the line is not supported between the antenna and the anchoring point.

LIGHTNING PROTECTION

Two or three types of lightning arresters for coaxial cable are available on the market. These are designed to be inserted between two lengths of coax cable. If the antenna feed point is at the top of a well-grounded tower, the arrester can be fastened securely to the top of the tower for grounding purposes. A short length of cable, terminated in a coaxial plug is then run from the antenna feed point to one receptacle of the arrester, while the transmission line is run from the other arrester receptacle to the station. Such arresters may also be placed at the entrance point to the station, if a suitable ground connection is available at that point (or arresters may be placed at both points for added insurance).

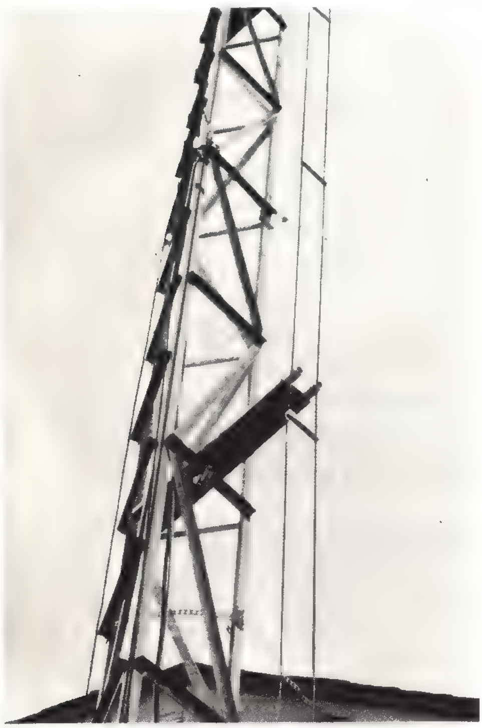


Fig. 11-10—A board fitted with standoff insulators, and clamped to the tower with U bolts keeps open-wire line suitably spaced from a tower. (W4NML).

The construction of a homemade arrester for open-wire line is shown in Fig. 11-14. This type of arrester can be adapted to ribbon line as shown in Fig. 11-15. The two TV standoff insulators should elevate the ribbon line an inch or so above the center member of the arrester. Sufficient insulation should be removed from the line where it crosses the arrester to permit soldering the arrester connecting leads.

Lightning Grounds

Lightning-ground connecting leads should be of conductor equivalent to at least No. 10 wire. The No. 8 aluminum wire used for TV-antenna grounds is satisfactory. Copper braid ¾-inch wide (Belden 8662-10) is also suitable. The conductor should run in a straight line to the grounding point. The ground connection may be made to a water-piping system, the grounded metal frame of a building, or to one or more ⅝-inch ground rods driven to a depth of at least 8 ft.

Supports For Wire Antennas

A prime consideration in the selection of a support for an antenna is that of structural safety. Building regulations in many localities require that a permit be secured in advance of the erection of structures of certain types, often including antenna poles or towers. In general, localities having such requirements also will have building safety codes that must be observed. Such regulations may govern the method and materials used in construction of, for example, a self-supporting tower. Checking with your local government building department before putting up a tower may save a good deal of difficulty later, since a tower would have to be taken down or modified if not approved by the building inspector on safety grounds.

Municipalities have the right and duty to enforce any reasonable regulations having to do with the safety of life or property. The courts generally have recognized, however, that municipal authority does not extend to esthetic questions; i.e., the fact that someone may object to the mere presence of a pole or tower, or an antenna structure, because in his opinion it detracts from the beauty of the neighborhood, is not grounds for refusing to issue a permit for a safe structure to be erected.

But, even where such regulations do not exist, or are not enforced, the amateur should be careful to select a type of support and a location for it that will minimize the chances of collapse and, if collapse does occur, will minimize the chances that someone will be injured or property damaged. A single injury can be far more costly than the price of a more rugged support.

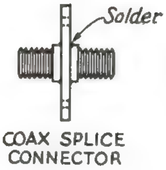


Fig. 11-13—Feedthrough connector for coax line. A connector of the type used to splice sections of coax line together is soldered into a hole cut in a brass mounting flange. Amphenol bulk adapter 83-F1 may be used instead.

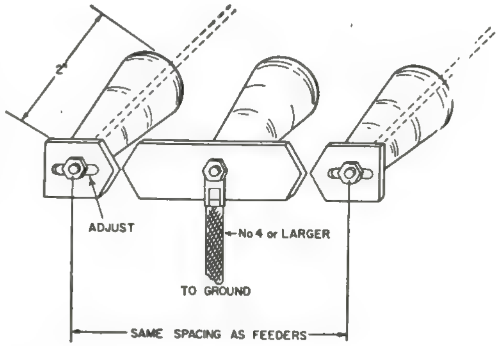


Fig. 11-14—A simple lightning arrester for open-wire line made from three standoff or feedthrough insulators and sections of ⅛ × ½-inch brass or copper strap. It should be installed in the line at the point where the line enters the station. The heavy ground lead should be as short and direct as possible. The gap setting should be adjusted to the minimum setting that will prohibit arcing when the transmitter is operated.

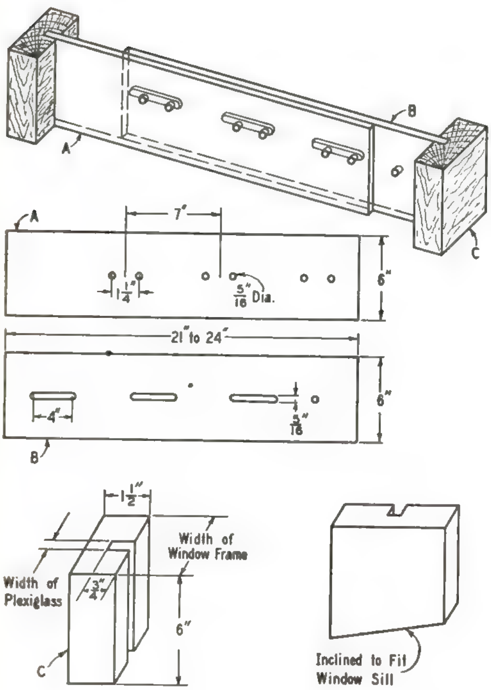


Fig. 11-12—An adjustable window lead-in panel made of two sheets of Lucite. A feedthrough connector for coax line can be made as shown in Fig. 11-13. Ceramic feedthrough insulators are suitable for open-wire line. (W1RVE).

TREES AS ANTENNA SUPPORTS

From the beginning of amateur radio, trees have been used widely for supporting wire antennas. Trees cost nothing, of course, and will often provide a means of supporting a wire antenna at considerable height. However, as an antenna support, a tree is highly unstable in the presence of wind, unless the tree is a very large one, and the antenna is suspended from

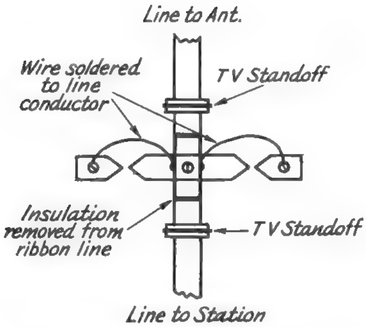


Fig. 11-15—The lightning arrester of Fig. 11-14 may be used with 300-ohm ribbon line in the manner shown in the sketch.

a point well down on the tree trunk. As a result, the antenna must be constructed much more sturdily than it would have to be with stable supports. Even with rugged construction, it is unlikely that an antenna suspended from a tree, or between trees, will stand up indefinitely, and occasional repair or replacement usually must be expected.

There are two general methods of securing a pulley to a tree. If the tree can be climbed safely to the desired level, a pulley can be wired to the trunk of the tree, as shown in Fig. 11-16. If, after passing the halyard through the pulley, both ends of the halyard are simply brought back down to ground along the trunk of the tree, there may be difficulty in bringing the antenna end of the halyard out where it will be clear of branches. To avoid this, one end of the halyard can be tied temporarily to the tree at the pulley level, while the remainder of the halyard is coiled up, and the coil thrown out horizontally from this level, in the direction in which the antenna will run. It may help to have the antenna end of the halyard weighted. Then, after attaching the antenna to the halyard, the other end is untied from the tree, passed through the pulley, and brought to ground along the tree trunk in as straight a line as possible. The halyard need be only long enough to reach the ground after the antenna has been hauled up, since additional rope can be tied to the halyard when it becomes necessary to lower the antenna.

The other method consists of passing a line over the tree from ground level, and using this line to haul a pulley up into the tree and hold it there. Several ingenious methods have been used to accomplish this. The simplest method employs a weighted pilot line, such as fishing line, or mason's chalk line. Grasping the line about two feet from the weight, the weight is swung back and forth, pendulum style, and then heaved with an underhand motion in the direction of the tree top. Several trials may be necessary to determine the optimum size of the weight for the line selected, the distance between the weight and the hand before throwing, and the point in the arc of the swing where the line is released. The weight, however, must be sufficiently large to assure that it will carry the pilot line back to ground after passing over the tree. Flipping the end of the line up and down so as to put a traveling wave on the line often helps to induce the weight to drop down if the weight is marginal. The higher the tree, the lighter the weight and the pilot line must be. A glove should be worn on the throwing hand, because a line running swiftly through the bare hand can cause a severe burn.

If there is a clear line of sight between ground and a particularly desirable crotch in the tree, it may be possible to hit the crotch eventually after a sufficient number of tries. Otherwise, it is best to try to heave the pilot line completely over the tree, as close to the

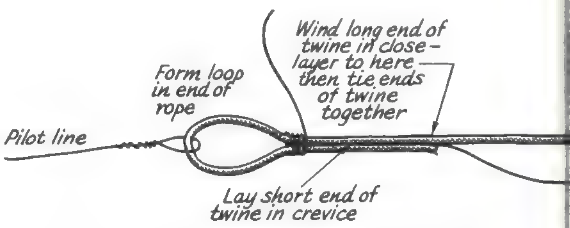


Fig. 11-17—In connecting the halyard to the pilot line, a large knot that might snag in the crotch of a tree should be avoided, as shown.

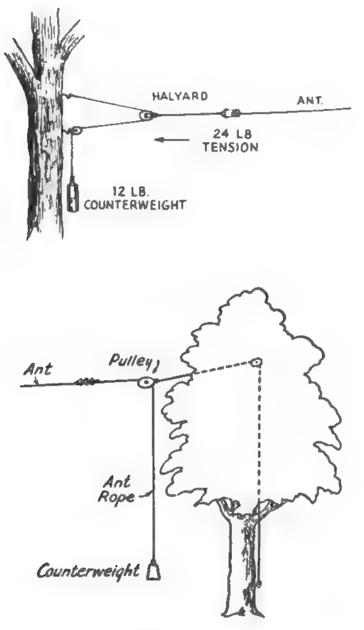


Fig. 11-16—Methods of counterweighting to minimize antenna movement. The first method (K2ZZF) limits the fall of the counterweight should the antenna break. It also has a 2 to 1 mechanical advantage, as indicated.

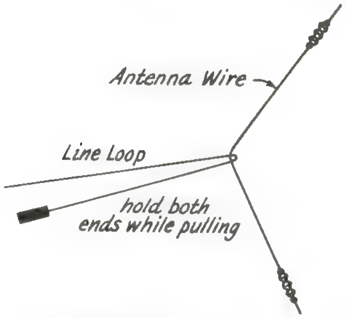


Fig. 11-18—A weighted line thrown over the antenna can be used to pull the antenna to one side to avoid overhanging obstructions, such as branches of trees in the path of the antenna, as the antenna is pulled up. When the obstruction has been cleared, the line can be removed by releasing one end.

center line of the tree as posible. If it is necessary to retrieve the line and start over again, the line should be drawn back very slowly, otherwise the swinging weight may wrap the line around a small limb, making retrieval impossible.

Stretching the line out in a straight line on the ground before throwing, may help to keep the line from snarling, but it places extra drag on the line, and the line may snag on obstructions overhanging the line, when it is thrown. Another method is to make a stationary reel by driving eight nails, arranged in a circle, through a 1-inch board. After winding the line around the circle formed by the nails, the line should reel off readily when the weighted end of the line is thrown. The board should be tilted at approximately right angles to the path of the throw.

Other devices that have been used successfully to pass a pilot line over a tree are the bow and arrow, with heavy thread tied to the arrow, and the short casting rod and spinning reel used by fishermen. Still another method that has been used where sufficient space is available is to fly a kite. After the kite has reached sufficient altitude, simply walk around the tree until the kite string lines up with the center of the tree. Then pay out string until the kite falls to earth. This method has been used successfully to pass a line over a patch of woods between two higher supports, which would have been impossible using any other method.

The pilot line can be used to pull successively heavier lines over the tree until one of adequate size to take the strain of the antenna has been reached. This line is then used to haul a pulley up into the tree after the antenna halyard has been threaded through the pulley. The line that holds the pulley must be capable of withstanding considerable chafing where it passes through the crotch, and at points where lower branches may rub against the standing part. For this reason, it may be advisable to use galvanized sash cord or stranded guy wire for raising the pulley.

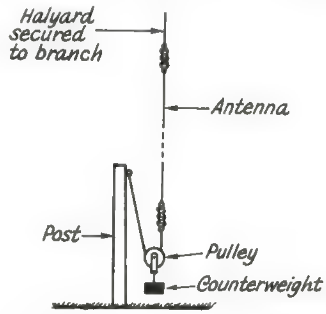


Fig. 11-20—Counterweight for a vertical antenna suspended from an overhanging tree branch.

Especially with larger sizes of line or cable, care must be taken when splicing the pilot line to the heavier line to use a splice that will minimize the chances that the splice cannot be coaxed through the tree crotch. One type of splice is shown in Fig. 11-17.

The crotch which the line first comes to rest in may not be sufficiently strong to stand up under the tension of the antenna. However, if the line has been passed over, or close to, the center line of the tree, it will usually break through the lighter crotches and finally come to rest in one sufficiently substantial lower down on the tree.

Needless to say, any of the suggested methods should be used with due respect to persons or property in the immediate vicinity. A child's sponge-rubber ball (baseball size) makes a safe weight for heaving a heavy thread line or fishing line.

If the antenna wire becomes snagged in lower branches of the tree when the wire is pulled up, or if branches of other trees in the vicinity interfere with raising the antenna, a

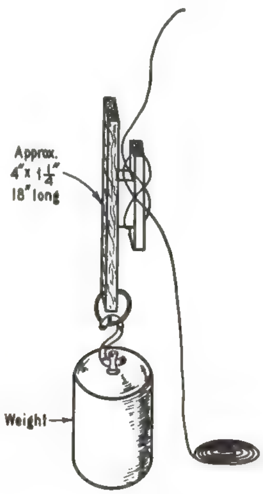


Fig. 11-19—The cleat avoids the necessity of having to untie a knot that may have been weather-hardened.

weighted line thrown over the antenna and slid along to the appropriate point is often helpful in pulling the antenna wire to one side to clear the interference as the antenna is being raised, as shown in Fig. 11-18.

Wind Compensation

The movement of an antenna suspended between supports that are not stable in wind can be reduced materially by the use of heavy springs, such as automotive valve springs, or by a counterweight at the end of one halyard, as shown in Fig. 11-16. The weight, which may be made up of junk-yard metal, window sash weights, or a galvanized pail filled with sand or stone, should be adjusted experimentally for best results under existing conditions. Fig. 11-19 shows a convenient way of fastening the counterweight to the halyard. It avoids the necessity for untying a knot in the halyard which may have hardened under tension and exposure to the weather.

Trees as Supports for Vertical Wire Antennas

Trees can often be used to support vertical as well as horizontal antennas. If the tree is a tall one with overhanging branches, the scheme of Fig. 11-20 may be used. The top end of the antenna is secured to a halyard passed over the limb, brought back to ground level, and fastened to the trunk of the tree. An earlier chapter describes a three-band ground-plane antenna that may be quite easily mounted in a tree.

MASTS

Where suitable trees are not available, or a more stable form of support is desired, masts are suitable for wire antennas of reasonable span length. At one time, most amateur masts were constructed of lumber, but the TV industry has brought out metal masting that is inexpensive, and much more durable than wood. However, there are one or two applications, such as supporting a vertical antenna (see Fig. 11-21) where wood is necessary. Two simple types of wood masts are shown in Figs. 11-22, 11-23, and 11-24.

The "A-Frame" Mast

A light and inexpensive mast is shown in Fig. 11-22. In lengths up to 40 feet it is very easy to erect and will stand without difficulty the pull of ordinary wire antenna systems. The lumber used is 2×2 straight-grained pine (which many lumber yards know as hemlock) or even fir stock. The uprights can be each as long as 22 feet (for a mast slightly over 40 feet high) and the crosspieces are cut to fit. Four pieces of 2×2, 22 feet long, will provide enough and to spare. The only other materials required are five ¼-inch carriage bolts 5½ inches

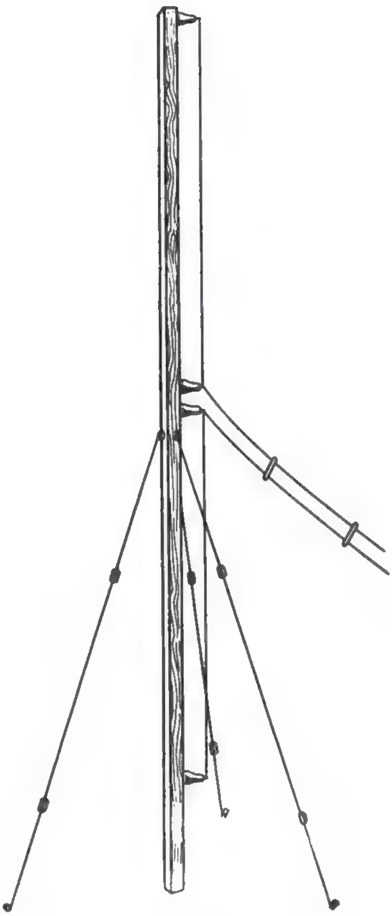


Fig. 11-21—A single 2 × 3 will serve as a mast for heights up to about 20 ft. This drawing shows how a vertical wire antenna may be installed on such a pole.

long, a few spikes, about 300 feet of No. 12 galvanized iron wire for the guys or stays, enough glazed-porcelain compression insulators ("eggs") to break up the guys into sections, and the usual pulley and halyard rope. If the strain insulators are put in every 20 feet approximately 15 of them will be enough.

After selecting and purchasing the lumber—which should be straight-grained and knot-free—three sawhorses or boxes should be set up and the mast assembled in the manner indicated in Fig. 11-23. At this stage it is a good plan to give the mast two coats of "outside-white" house paint or latex.

After the second coat of paint is dry, attach the guys and rig the pulley for the antenna halyard. The pulley anchorage should be at the point where the top stays are attached so that the back stay will assume the greater part of the load tension. It is better to use wire wrapping around the stick, with a small through-bolt to prevent sliding down, than to use eye bolts.

If the mast is to stand on the ground, a couple of stakes should be driven to keep the bottom from slipping. At this point the mast may be "walked up" by a pair of helpers. If it is to go on a roof, first stand it up against the side of the building and then hoist it, from the roof, keeping it vertical. The whole assembly is light enough for two men to perform the complete operation—lifting the mast, carrying it to its permanent berth, and fastening the guys—with the mast vertical all the while. It is therefore entirely practicable to put up this kind of mast on a small flat area of roof that would prohibit the erection of one that had to be raised vertical in its final location.

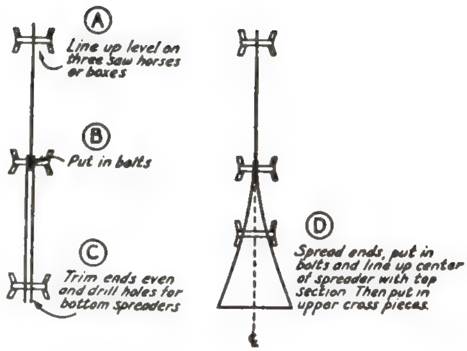


Fig. 11-23—Method of assembling the "A-frame" mast.

A Wood Tilt-Over Mast

Fig. 11-24 shows the construction of a wood mast that can be tilted over it if it become necessary to replace the pulley or halyards. Without the tension of the antenna, it can be raised quite easily by pulling down on the raising rope. After the upper part has been raised, two bolts should be passed through the upper part of the lower section and the lower part of the upper section to lock the upper section in place. Then the stiffening block is inserted in the lower section, and clamped tight with bolts, as indicated. For long life, all parts of the mast should be given two coats of house paint or latex, and the bolts sprayed with acrylic. The base block should be coated with creosote, or smeared with roofing compound before burying. Guy wires should be arranged as shown in Fig. 11-22.

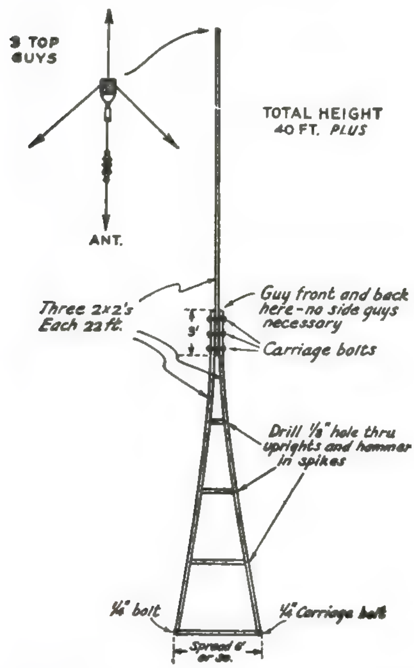


Fig. 11-22—The "A-frame" mast, lightweight and easily constructed and erected.

TV Masting

TV masting is available in 5- and 10-ft. lengths, 1¼ inches in diameter, in both steel and aluminum. These sections are crimped at one end to permit sections to be joined together. However, a form that will usually be found more convenient is the telescoping TV mast available from many electronic supply houses. These masts may be obtained with 3, 4, or 5 10-ft. sections, and come complete with guying rings, and a means of locking the sections in place after they have been extended. These masts are stronger than the nontelescoping type because the top section is 1¼ inches in diameter, and the diameter increases toward the bottom section which is 2½ inches in diameter in the 50-ft. mast.

Guy rings are provided at 10-ft. intervals, but guys may not be required at all points. Guys at the top are essential, and at least one other set near the center of the mast will usually be found necessary to keep the mast from bowing. If the mast has any tendency to whip in the wind, or to bow under the stress of the antenna, additional guys should be added at the obvious points.

MAST GUYING

Three guy wires in each set will usually be adequate for a mast. These should be spaced equally around the mast. The number of sets of guys will depend on the height of the mast, its natural sturdiness, and the required antenna tension. A 30-ft. mast will usually require two sets of guys, while a 50-ft. mast will need at least three sets. One guy of the top set should be run in a direction directly opposite to the direction in which the antenna will run, the other two being spaced 120 degrees in respect to the first, as shown in Fig. 11-22.

The general rule is that the top guys should be anchored at distances from the base of the mast equal to not less than 60% of the height of the mast. At the 60% distance, the stress on the guy wire opposite the antenna will be approximately twice the tension on the antenna.

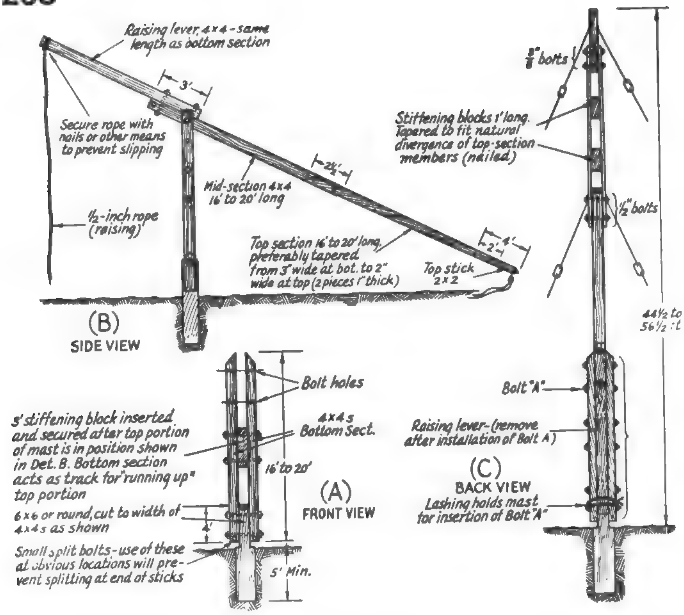


Fig. 11-24—A wooden tilt-over mast made principally of 4 × 4-inch lumber.

As the distance between the guy anchorage and the base of the mast is decreased, the tension on the rear guy in proportion to the tension on the antenna rises rapidly, the extra tension resulting in additional compression on the mast, increasing the tendency for the mast to buckle.

The function of additional sets of guys is to correct for any tendency that the mast may have to bow or buckle under the compression imposed by the top guys. To avoid possible mechanical resonance in the mast that might cause the mast to have a tendency to vibrate, the sets of guys should not be spaced equally on the mast. A second set of guys should be placed at approximately 60% of the distance between the ground and the top of the mast, while a third set should be placed at about 60% of the distance between the ground and the second set.

The additional sets of guys should be anchored at distances from the base of the mast equal to not less than 60% of the distance between ground and the points of attachment on the mast. In practice, the same anchors are usually used for all sets of guys, which means that the latter requirement is met automatically, if the top set has been anchored at the correct distance.

To avoid electrical resonances which might cause distortion of the normal radiation pattern of the antena, it is advisable to break each guy into sections of 19 to 20 ft. by the insertion of strain insulators (see Figs. 11-25 and 11-26).

Guy Material

Within their stress ratings, any of the halyard materials listed in Table 11-II may be used for the construction of guys. The nonmetalic materials have the advantage that they do not have to be broken up into sections to avoid resonances, but all of these materials are subject to stretching, which may cause mechanical problems in permanent installations. At rated working load tension, dry manila rope stretches about 5 percent, while nylon rope stretches about 20 percent.

The antenna wire listed in Table 11-I is also suitable for guys, particularly the copper-clad steel types. Solid galvanized steel wire is also used widely for making guys. This wire has approximately twice the tension ratings of similar sizes of copper-clad wire, but it is more susceptible to corrosion.

Guy Anchors

Figs. 11-27 and 11-28 show two different styles of guy anchors. In Fig. 11-27, one or more pipes are driven into the earth at right angles to the guy wire. If a single pipe proves to be inadequate, another pipe can be added in tandem, as shown. Steel fence posts may be used in the same manner. Fig. 11-28 shows a "dead-man" type of anchor. The buried anchor may consist of one or more pipes 5 or 6 ft. long, or scrap automobile parts, such as bumpers or

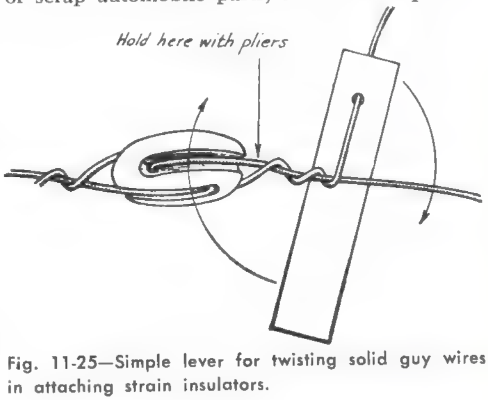


Fig. 11-25—Simple lever for twisting solid guy wires in attaching strain insulators.

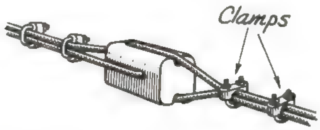


Fig. 11-26—Stranded guy wire should be attached to strain insulators by means of standard cable clamps to fit the size of wire used.

wheels. The anchors should be buried 3 or 4 ft. in the ground. Some tower manufacturers make heavy auger-type anchors that screw into the earth. These anchors are usually heavier than required for guying a mast, although they may be more convenient to install. Trees and buildings may also be used as guy anchorages if they are located appropriately. Care should be exercised, however, to make sure that the tree is of adequate size, or that the fastening to a building can be made sufficiently secure.

Guy Adjustment

Most troubles that are encountered in mast guying are a result of pulling the guy wires too tight. Guy-wire tension should never be more than is necessary to correct for obvious bowing or movement under wind pressure. In most cases, the tension needed will not require the use of turnbuckles, with the possible exception of the guy opposite the antenna. If any great difficulty is experienced in eliminating bowing from the mast, the antenna tension should be reduced.

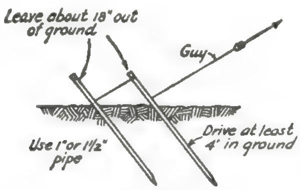


Fig. 11-27—Driven guy anchors. One pipe will usually be sufficient for a small mast. For added strength, a second pipe may be added, as shown.

ERECTING A MAST

The erection of a mast of 30 ft. or less can usually be done by simply "walking" the mast up after blocking the bottom end securely so that it can neither slip along the ground or upend when the mast is raised. A man should be stationed at each guy wire, and in the last stages of raising, some assistance may be desirable by pulling on the proper guy wire. Further assistance may be gained by using the halyards in the same manner. As the mast is raised, it may be helpful to follow the under side of the mast with a scissors rest (Fig. 11-29), should a pause in the hoisting become necessary. The rest may also be used to assist in the raising, if a man is used on each leg.

As the mast nears the vertical position, those holding the guy wires should be ready to make the guys fast temporarily to prevent the mast from falling in one direction or another. The guys can then be adjusted, one at a time, until the mast is perfectly straight.

For a mast of over 30 ft., a "gin" of some form may be required, as shown in Fig. 11-29. Several turns of rope are wound around a point on the mast above center. The ends of the rope are then brought together and passed over the limb of a tree. The rope should be pulled as the mast is walked up to keep the mast from bending at the center. If a tree is not available, a

Fig. 11-28—Buried "dead-man" type of guy anchor (see text).

post, such as a 2×4, temporarily erected and guyed, can be used. After the mast has been erected, the assisting rope can be removed by walking one end around the mast (inside the guy wires).

Other Supports

Much sturdier supports are telephone poles and towers. These types of supports are discussed later in reference to rotatable antennas. Such supports may require no guying, but they are not often used solely for the support of wire antennas because of their relative high cost. However, for antenna heights in excess of 50 ft., they are usually the most practical form of support.

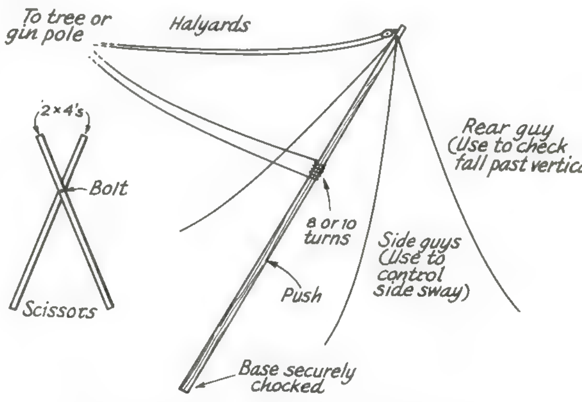


Fig. 11-29—Pulling on a gin line fastened slightly above the center point of the mast, and on the halyards will assist materially in erecting a tall mast. The tensions should be such as to keep the mast in as straight a line as possible. The scissors may be used to push on the under side and to serve as a rest in case a pause in raising becomes necessary.

Construction of Rotatable Antennas

Yagi Antennas

A Yagi-type antenna for the high-frequency bands is usually an assembly of metallic tubing. The chief constructional problems are in providing devices for connecting the various tubular members together to form a satisfactory mechanical structure.

MATERIALS

Noncorrosive materials should be used as far as possible, not only to assure good electrical contact between members, but also to prevent mechanical failure. Unprotected ferrous materials rust very rapidly when exposed to weather, and it soon becomes impossible to tighten bolts, or to remove them should this become desirable, without the use of a hacksaw.

Aluminum alloy is most commonly used for the tubular members, not only because it resists corrosion, but also because of its light weight.

Other hardware, such as bolts and nuts, U bolts, and clamps should be aluminum (if strong enough for the application) or zinc-coated-galvanized, chrome plated, or stainless steel. Stainless-steel gear-type hose clamps are available at most gasoline stations. Most of the bright U bolts sold in hardware stores appear to resist corrosion quite well. Copper or copper alloys, such as brass, should not be used in contact with aluminum. It is highly important that lockwashers be used under all assembly nuts or screws to prevent loosening. To resist both corrosion and loosening, bolts and nuts should be sprayed with acrylic after tightening.

BOOMS

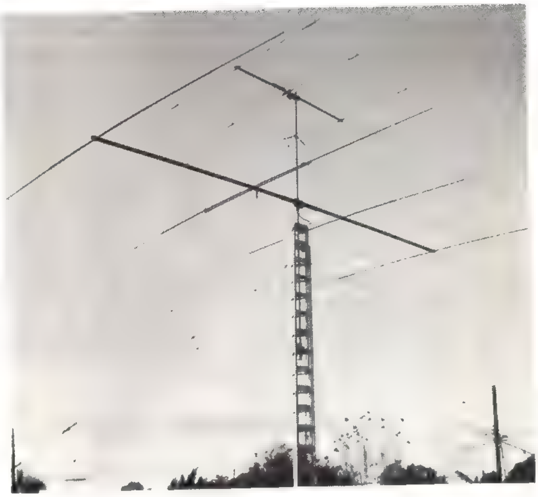
The diameter of the tubing to be chosen for making the boom and elements depends upon the length of the members, the wall thickness of the pipe or tubing selected, and the type of alloy. Alloy types most suitable for antenna construction are those designated 6063-T6, 6061-T6 and 2024-T4. Type 6061-T6 probably has the widest distribution. As an approximate guide, the minimum dimensions for booms carrying up to four elements are as follows:

Boom Length	*Pipe Size*	*O.D.*	*Wall*
10-15 ft.	1"	1.315"	0.135"
15-20 ft.	1½"	1.9"	0.145"
20-25 ft.	2"	2.375"	0.154"

Pipe, rather than tubing, is listed for boom use because the standard length of tubing is 12 ft., whereas pipe can be obtained in 20-ft. as well as 12-ft. lengths.

Mounting the Boom

The chief problem in fastening the boom to the rotating mast is to prevent slippage between



Typical Yagi antennas (VE2AES/W6).

the two. Fig. 12-1 shows one method of attaching a one-piece boom to a mast. It is suitable principally for relatively short booms. The rated maximum wind loading on some free-standing towers is quite small, and if the plate is made large enough to provide adequate support for a longer boom, its area may become a significant part of the wind load.

The method of mounting shown in Fig. 12-2 results in minimum wind loading, and is more suitable for longer booms, since the plate size can be increased without increasing the wind load significantly. For booms of 20 to 25 ft., the plate should be ½ inch thick, and about 2 ft. long. A ¼-inch plate might be used if it is stiffened by bolting iron or heavy-aluminum angle stock along the edges parallel to the boom.

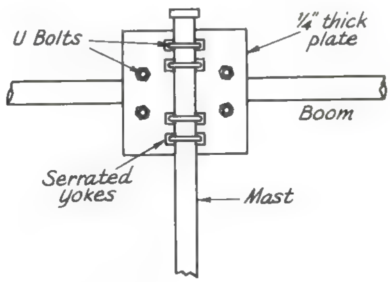


Fig. 12-1—A method of mounting a one-piece boom less than 15 ft. long. The plate should be at least ¼ inch thick, and 10 to 12 inches square. Serrated yokes between the plate and the mast reduce the tendency for the boom to slip on the mast. This type of junction is also suitable for element-to-boom mounting.

A third method of mounting a boom is shown in Fig. 12-3. This consists of a sleeve of pipe welded across the end of the mast. The boom is inserted inside the sleeve. This method is suitable for mounting two-piece as well as single-piece booms. A two-piece boom may be desirable where boom material of sufficient length for a one-piece boom cannot be obtained. For a one-piece boom, the sleeve should be 12 to 18 inches long. For a two-piece boom, it should be at least 24 inches long. Use a 2½-inch pipe-size sleeve for a 1½-inch pipe-size boom, a 2-inch sleeve for a 1½-inch pipe boom, and a 1¼-inch sleeve for a 1-inch boom.

The pipe and boom should be drilled simultaneously after inserting the boom in proper position. A method of drilling accurate holes in aluminum tubing is shown in Fig. 12-4. After drilling, spacers of suitable thickness to make a reasonably close fit should be made and shaped to the contours of the boom and sleeve. To keep the spacers in place while the boom is inserted and fastened with stainless-steel bolts, the spacers can be cemented over the holes in the boom.

The mounting methods of Figs. 12-2 and 12-3 are particularly desirable, since they eliminate any possibility of slippage between the boom and the mast.

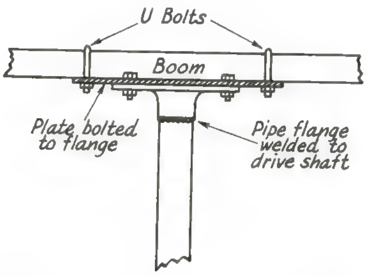


Fig. 12-2—A method of boom mounting that minimizes wind loading and eliminates slippage between boom and mast.

MAKING YAGI ELEMENTS

Elements for 10-, 15-, or even 20-meter antennas can be made in one or two sections without the necessity for telescoping if material in suitable lengths can be obtained. The minimum tubing sizes to be used in this case are approximately as follows:

Band	*O.D.*	*Wall*
10 m.	¾″	0.058″
15 m.	1″	0.058″
20 m.	1½″	0.125″
20 m.	2″	0.083″

However, to reduce the weight and wind resistance, it is usually desirable to reduce the diameter toward the outer ends of the elements. In doing this, it may also be possible to reduce the diameter and/or the wall thickness of the

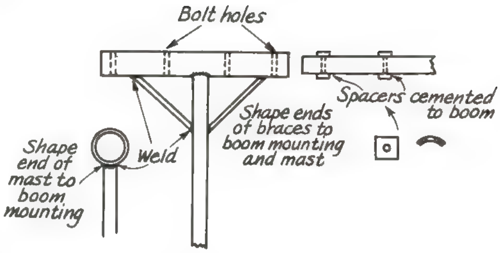


Fig. 12-3—Sleeve-type boom mounting suitable for either one- or two-piece booms.

material, particularly in the case of 20-meter elements.

Telescoping

Standard sizes of aluminum tubing are not graduated in steps that are most desirable for making telescoped joints between sections of different diameters. Outside diameters are available in ⅛-inch steps down to ½ inch, and in 1⁄16-inch steps for smaller sizes. Tubing of each diameter is available in several different wall thicknesses, the wall thickness determining the inner diameter. Adjacent diameter sizes in the series will telescope if the wall thickness is 0.047 inch, but this wall thickness is rather light, although it might be suitable for 10-meter elements if it can be obtained. However, it is more desirable to change diameter in steps greater than ⅛ inch so that more weight and area may be saved. Tubing differing in outside diameter by ¼ inch will telescope if the wall thickness is 0.109. But this wall is heavier than is usually necessary and also may not be widely available.

With either of the above combinations, the sections can be clamped together by slitting the end of the larger-diameter member to a depth of about 3 inches by cutting diametrically across the end of the tubing with a hacksaw. The smaller tubing should be inserted 6 or 8 inches into the larger, and the joint secured by a gear-type stainless-steel hose clamp around the outer member, about ½ inch from its end. If there is any tendency for sideway or up-and-down movement, it can be eliminated by denting the outer tubing slightly at intervals of 90 degrees with a nail set or screwdriver blade, about 1 inch beyond the end of the slit, as shown in Fig. 12-5A.

A commonly-available wall thickness is 0.58 inch. Tubing having this wall thickness and having a difference in outside diameter of ¼ inch can be telescoped very satisfactorily by the method shown in Fig. 12-5B. The shims are made by slicing a 4-inch length of the smaller-diameter tubing lengthwise. Then clamp the two shims and the end of the tubing from which the shims have been cut in a vise to force the shims over the tubing. Then squeeze at other points as necessary to make the shims conform as closely as possible to the contour of the tubing.

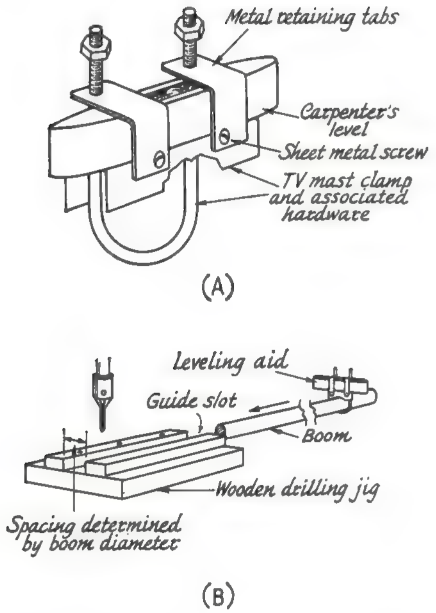


Fig. 12-4—A method of drilling accurate holes in metallic tubing on a drill press. After clamping the wood jig central on the drill-press table, the tubing is held in a position that keeps the level bubble centered. *(W5UQR)*

In assembly of the two tubing sections, the shims should be placed diametrically opposite on the smaller tubing, and oriented so that they will be centered on the slits in the larger tubing. Place a mark on the small tubing at approximately 4 inches from the end that will be inserted. Keep this line in sight during the insertion. During the last stage of the insertion, the shims can be nudged in against a sharp corner of the vise until the ends of the shims are flush with the end of the outer tubing. Then the inner tubing can be tapped in to bring the marked line to the same point. The joint can then be secured by tightening the clamp.

Fig. 12-6 indicates typical sets of dimensions for telescoped-joint elements. Overall lengths may be determined by reference to the charts of Fig. 12-7.

Element-to-Boom Mounting

The general method shown in Fig. 12-1 can be used to fasten one-piece (not broken at center) elements to the boom, with the boom replacing the mast, and the element replacing the boom in that sketch. A single pair of U bolts around the boom will usually be sufficient to keep the elements from twisting on the boom. A plate 12 inches square and ¼ inch thick will usually be the maximum required.

Fig. 12-8A shows a method of mounting a two-piece driven element. The channel should be 2 to 3 ft. long, depending on the length of the element. An insulator spacing of about 12 inches is recommended for a 20-meter element. The insulators are 1¾ inches high with a base 2½ inches in diameter (Johnson type 135-53) with metal mounting flanges (Johnson 135-90). The bolts are ¼-inch stainless steel. If the element has a tendency to twist on the boom with the single U-bolt clamp, a ¼-inch plate, 1½ inches wider than the U bolt, and running 7 or 8 inches along the boom, can be placed under the channel, and a second U bolt used at the other end of the plate, as shown in Fig. 12-8B.

Element Adjustment

The lengths of the elements in a two- or three-element Yagi do not vary greatly with the element spacing, but the gain of the array, the bandwidth and the impedance of the driven element all do. It is advisable to make the element spacing in a 3-element beam as great as possible (up to 0.2 wavelength), and to provide for adjusting the matching of the antenna to the feeder after the beam is installed on its support. A 2-element beam should have a spacing of 0.1 to 0.15 wavelength, depending on whether the parasitic element is to be used as a director or reflector. The parasitic director spaced 0.1 theoretically gives slightly more gain than the parasitic reflector spaced 0.15 wavelength. Gain variation with spacing in a 3-element beam is treated in Chapter Four.

Element lengths for any of the spacings mentioned in Chapter Four can be found in the charts of Fig. 12-7. If the lengths are taken from the graphs, very little improvement in forward gain will be obtained by adjustment of the elements after the antenna system is in place, but the front-to-back ratio can be increased by careful tuning of the reflector.

GAMMA MATCH CONSTRUCTION

Figs. 12-9, 12-10 and 12-11 show various views of mounting gamma-match components

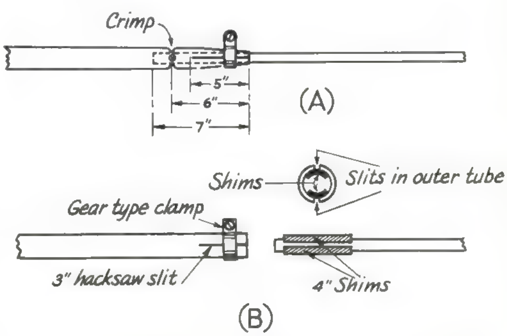


Fig. 12-5—Methods of joining tubing sections in making tapered Yagi elements. The simple method of A can be used with close-fitting sections if the material can be found. However, it provides only a small amount of tapering unless many sections, or sections with heavy walls are used. The method shown at B provides good tapering with 17-gauge tubing. Both methods are discussed in the text.

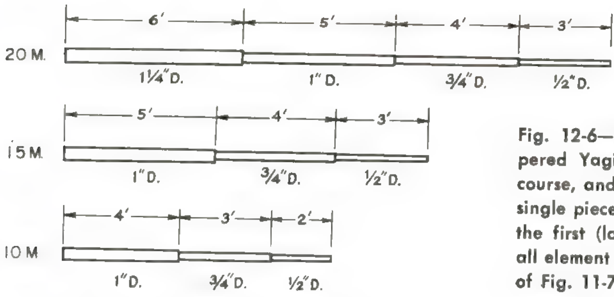


Fig. 12-6—Suggested proportions for one side of tapered Yagi elements. The other side is identical, of course, and the center section of the element can be a single piece twice as long as the length shown here for the first (largest diameter) section. Appropriate over-all element lengths may be determined from the graphs of Fig. 11-7.

on a driven element. In Fig. 12-9, a bracket fastened to the boom/element assembly plate under one of the U-bolt nuts, serves as a mounting for a tin can of suitable dimensions to house the gamma capacitor. A coffee or shortening can with a snap-on plastic cover is most suitable. The capacitor is mounted on a ¼-inch Lucite disk made to fit the can closely. The disk is fastened in place with small screws tapped into the edge of the Lucite, or by drilling tiny press-fit holes for brads and cementing them in.

The stator connecting lead should be made long enough to assure that it will reach the connector clip on the gamma rod, and be passed through a hole in the disk prior to assembly. The rotor connecting lead should be made of insulated wire, and should be made long enough so that it can be soldered externally to the coaxial connector before mounting the latter.

In assembly, the can is mounted on the bracket first. Then the capacitor assembly is inserted in the can with the rotor lead passing out through the connector hole. The Lucite disk is then fastened in place, the rotor lead soldered to the connector with the shortest lead possible, and the connector mounted with screws tapped into the bracket.

A small hole should be drilled in the cover of the can so that the capacitor may be adjusted with a screwdriver. This hole can be covered with waterproof adhesive tape when the adjustment is complete. A ¼-inch hole should be cut on the under side of the can to discourage condensation.

Fig. 12-11 shows the details for mounting the gamma rod. Suitable lengths and spacings for this rod are shown in the table of Fig. 12-12.

Fig. 12-12 shows the construction of a gamma

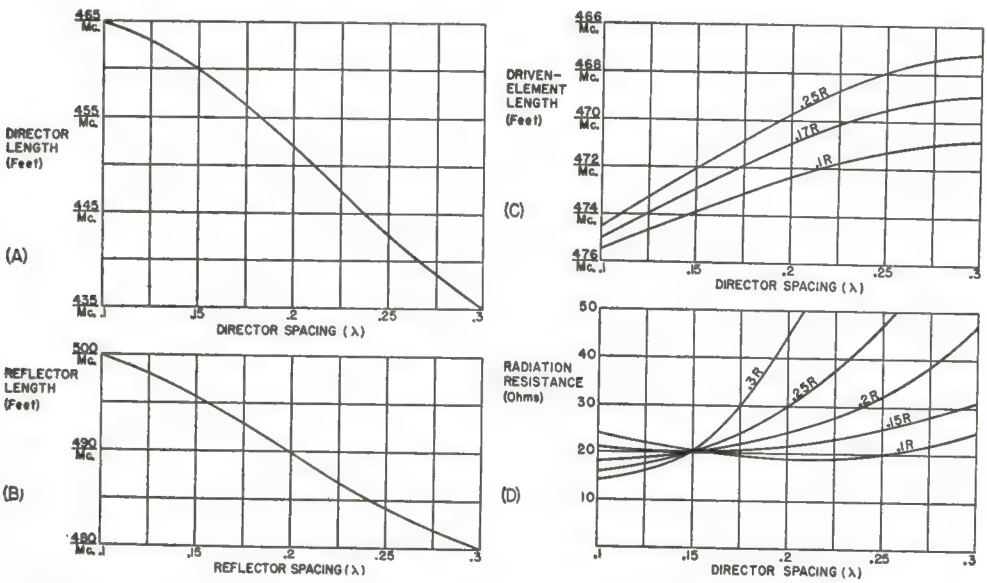


Fig. 12-7—Element lengths for 3-element Yagis. These lengths will hold closely for tubing elements supported at or near the center. The radiation resistance (D) is useful information in planning for a matching system, but it is subject to variation with height above ground and must be considered an approximation.

The driven-element length (C) may require modification for tuning out reactance if a gamma-match feed system is used.

A 0.2D-0.2R beam cut for 28.6 Mc. would have a director length of 452/28.6 = 15.8 = 15 feet 10 inches, a reflector length of 490/28.6 = 17.1 = 17 feet 1 inch, and a driven-element length of 470.5/28.6 = 16.45 = 16 feet 5 inches.

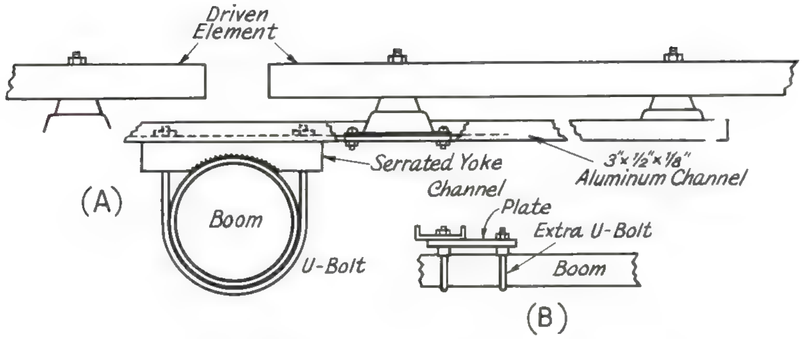


Fig. 12-8—Method of mounting two-piece driven elements.

matching section in which the capacitor and rod are combined as a tubular unit. The capacitor consists of the ½-inch aluminum tubing coaxial with the ¾ inch aluminum tubing, with the polystyrene tubing between acting as a dielectric. The capacitance is approximately 15 pf. per inch of engaged aluminum tubing. Dimension *C* is the length of the gamma rod, which is an extension of the ½-inch tubing. The table in the drawing gives approximate dimensions, but these have to be varied to effect an actual match to the coax cable.

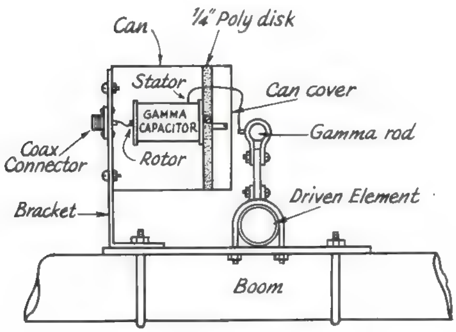


Fig. 12-9—Method of mounting a gamma capacitor.

The adjustment process using this capacitor is as follows: First take a trial setting of the supporting strap and measure the standing-wave ratio at the desired center frequency, with the gamma rod inserted all the way in the capacitor. Then adjust the strap position for minimum s.w.r. and clamp the strap firmly on the driven element at this position. Next, slide the gamma rod through the capacitor and strap until the setting is found that gives the lowest possible s.w.r. If this is not 1 to 1, leave the capacitor at that setting and readjust the strap for minimum, then repeat the capacitor adjustment. Continue until the lowest possible s.w.r. is obtained.

MINIATURE YAGI ANTENNAS

A 20-meter Yagi no larger than the usual 10-meter beam can be made by using center-loaded elements and close spacing. Such an antenna will show good directivity and can be rotated with a TV-antenna rotator.

Constructional details of a typical antenna of this type are shown in Figs. 12-13 and 12-14. The loading coils are space-wound by inter-winding plumb line (sometimes known as chalk line) with the No. 12 wire coils. The coil ends are secured by drilling small holes through the polystyrene bar, as shown in Fig. 12-14. The coils should be sprayed or painted with Krylon before installing the protective Lucite tubes.

The beam will require 4-foot lengths of the tubings indicated in Fig. 12-13A. For good telescoping, element wall thickness of 0.058 inch is recommended. The ends of the tubing sections should be slotted to permit adjustment, and secured with clamps so that the joints will not work loose in the wind. Perforated ground clamps can be used for this purpose. The boom is a 12-foot length of 1½-inch o.d. 6061-T6 aluminum tubing, with 0.125-inch wall. Fig. 12-15 shows how the boom is mounted on the pipe mast.

The line is coupled and matched at the center

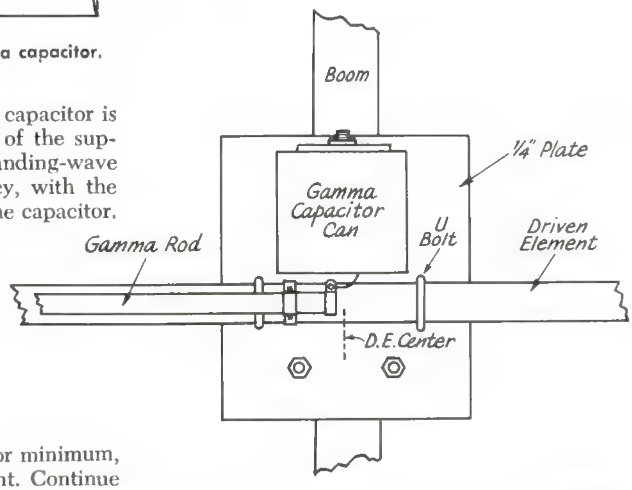


Fig. 12-10—Top view of gamma-capacitor assembly.

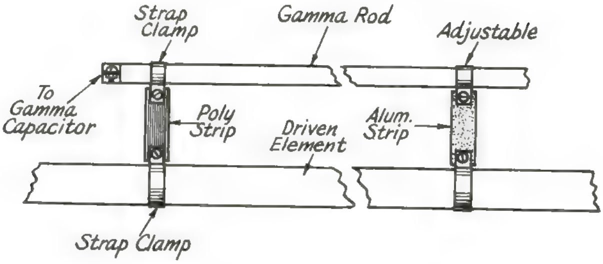


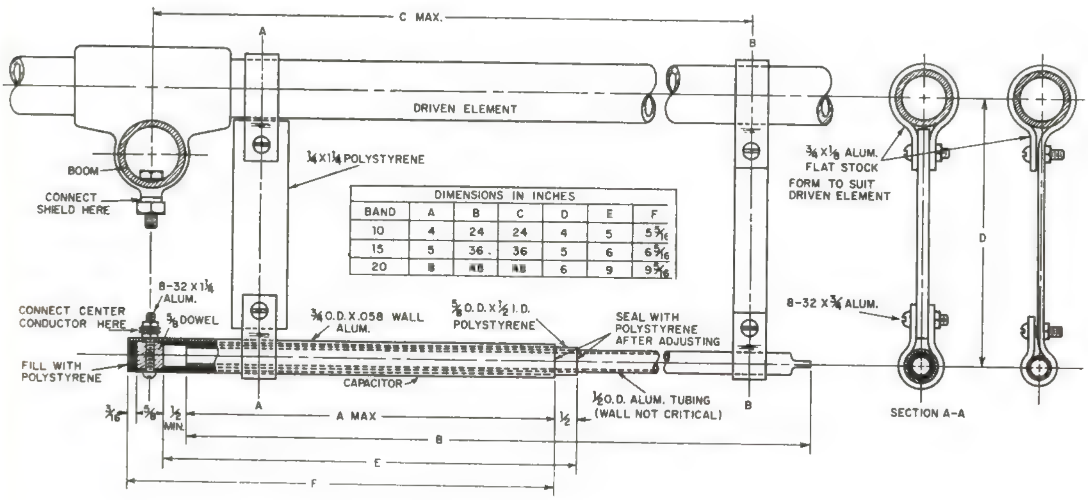
Fig. 12-11—Method of mounting gamma rod. See table of Fig. 12-12 for dimensions.

of the driven element through adjustment of the link wound on the outside of the Lucite tubing. To check the adjustment of the elements, first resonate the driven element to the desired frequency in the 14-Mc. band with a grid-dip oscillator. Then resonate the director to approximately 14.8 Mc., and the reflector to approximately 13.6 Mc. This is not critical and only serves as a rough point for the final tuning, which is done by use of a conventional field-strength indicator. Check the transmitter loading and readjust if necessary. Adjust the director for maximum forward gain, and then adjust the reflector for maximum forward gain. At this point, check the driven element for resonance and readjust if necessary. Turn the reflector toward the field-strength indicator and adjust for back cut off. This must be done in small steps. Do not expect the attenuation off the sides of a short beam to be as high as that obtained with full-length elements. The s.w.r. of the line feeding the antenna can be checked with a bridge, and after the elements have been tuned, a final adjustment of the s.w.r. can be made by adjusting the coupling at the antenna by varying the loading-coil turns and spacing. As in any beam, the s.w.r. will depend upon this adjustment and not on any that can be made at the transmitter. Transmitter coupling is the usual for any coaxial line.

TRIBAND TRAP YAGI ANTENNAS

Probably the most-used type of multiband system at the present time is the one using "traps" in the elements to give multiple resonances in two or more amateur bands. The general principles of multiband dipoles (each element of a parasitic Yagi is a dipole, of course) have been discussed in Chapter Six. These same principles apply in parasitic beams, with due allowance for the fact that the driven element, reflector and director are all tuned to slightly different frequencies in order to secure a desired directional pattern.

Typically, a trap Yagi will be either a two- or three-element affair with traps placed and tuned appropriately to result in the desired resonances in the selected bands. The schematic circuit in Fig. 12-16 is the electrical equivalent of a three-element three-band antenna using a



DIMENSIONS IN INCHES						
BAND	A	B	C	D	E	F
10	4	24	24	4	5	5 5/16
15	5	36	36	5	6	6 5/16
20	8	48	48	6	9	9 5/16

Fig. 12-12—Constructional details of a gamma matching section for 52-ohm coax line. The reactance-compensating capacitor is in tubular form. It is made by dividing the gamma rod or bar into two telescoping sections separated by a length of polystyrene tubing which serves as the dielectric. *(W2VS)*

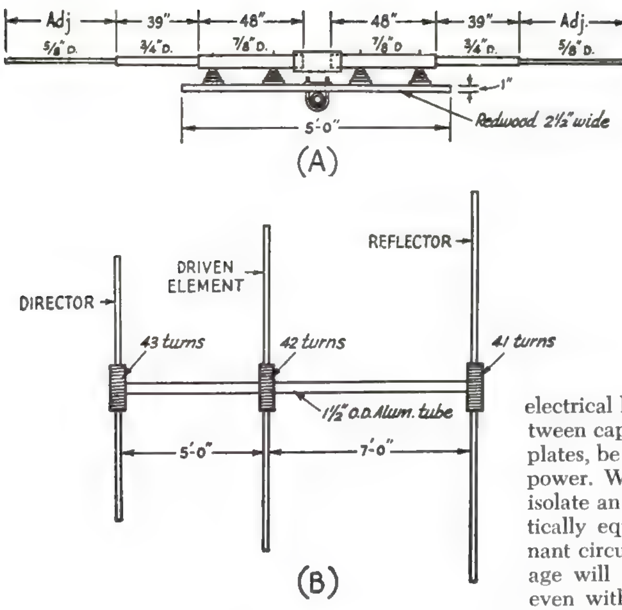


Fig. 12-13—Dimensions of a compact 14-Mc. Yagi antenna (WØVZC-WØQFG). A — Side view of a typical-element. TV-antenna "U" clamps hold the support arms to the boom. Birnbach 4176 insulators support the elements (or see Fig. 12-8). B—Top plan of the beam showing element spacing and loading-coil dimensions. Elements are made of aluminum tubing. Construction of the loading coils and adjustment of the elements are discussed in the text. End-section lengths of 41 inches for the reflector, 40 inches for the driven element, and 10 inches for the director will be close to optimum.

driven element, reflector and director. The effect of the traps is to divorce the ends of the elements from the central sections at the higher frequencies. For example, in Fig. 12-16 length *A* would be active at 28 Mc., length *B* at 21 Mc., and the entire antenna at 14 Mc. Because the outer sections, at least, are shorter than a normal unloaded element this system does not have as great a band width as a full-size single-band antenna. The actual reduction in band width depends on the element length, which can be controlled to a considerable extent by the design of the traps. Arriving at suitable values for trap constants, as well as the lengths of the linear segments of the elements, is almost wholly a matter of cut-and-try.

Trap Construction

The *LC* circuits used as isolating and loading elements in a multiband system can take any convenient form, the important mechanical requirements being light weight and adequate weatherproofing. It is also necessary that the electrical losses be low, and that the spacing between capacitor plates, whatever the form of the plates, be great enough to handle the transmitter power. When the circuit is acting as a trap to isolate an outer section of the antenna it is practically equivalent to an unloaded parallel-resonant circuit. This means that a rather high voltage will appear across the ends of the circuit even with moderate amounts of power.

A form of construction that has had considerable use with the tubing elements of rotatable beams uses a capacitor consisting of two concentric metal cylinders of approximately the same size as the tubing used for the antenna elements. These two "plates" (forming a cylindrical capacitor) are usually separated by low-loss insulating tubing—such as polystyrene—to make a mechanically stable assembly and to increase the breakdown voltage between the plates. The inductor is made with spaced turns of heavy wire or light metal tubing with the coil having a larger diameter than the capacitor (it should be several times the capacitor diameter, for minimizing losses) so it can be mounted coaxially with the capacitor. This forms a single assembly that can be inserted between sections of the beam element.

Fig. 12-17 gives the details of a representative trap of this type. This particular trap was designed for use with half-inch diameter tubing elements of a conventional 10-meter parasitic-type beam to extend the frequency coverage to include the 21-Mc. band. The following description of its construction can easily be modi-

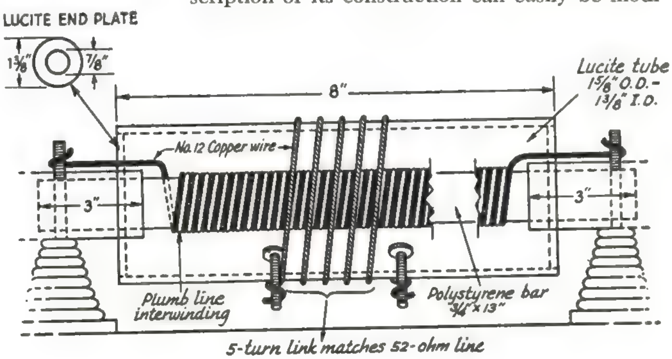


Fig. 12-14—Detailed sketch of the loading and coupling coils at the center of the driven element, and its mounting. Similar loading coils (see text) are used at the centers of the director and reflector.

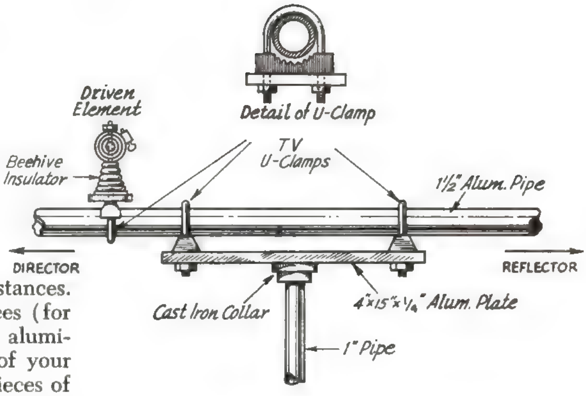


Fig. 12-15—The boom of the miniature Yagi is mounted on a rectangular aluminum plate fastened to the top of the pipe mast. A pipe flange can be substituted for the collar shown. This sketch also shows an end view of an element mounting.

fied as required by individual circumstances.

First cut to a length of 6 inches 6 pieces (for a 3-element beam—2 to each element) of aluminum tubing that will fit over the ends of your elements. Then cut to 3-inch lengths 6 pieces of polystyrene tubing that will fit inside the aluminum pieces. Next, cut to 6-inch lengths 6 pieces of aluminum tubing that will fit inside the poly

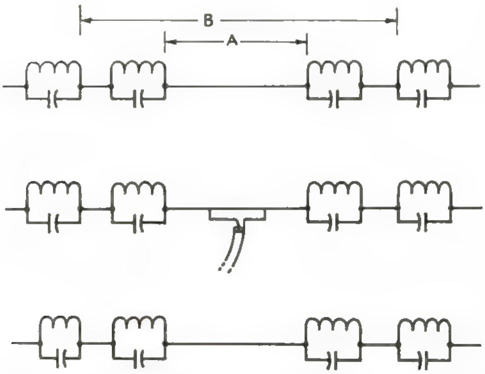


Fig. 12-16—Electrical diagram of a 3-band 3-element Yagi. The tuned circuits in the elements divorce the central sections from the outer sections on the higher frequencies.

tubing. These pieces of tubing form the trap capacitor.

Before assembling them, cut a 1-inch saw slot in one end of each of the aluminum pieces, and near the opposite end of each piece drill small holes approximately as shown in the sketch. Do not drill the poly tubing.

Insert the poly tubing into the larger aluminum tubing to a depth of 2½ inches, and then insert the smaller aluminum tubing into the poly tubing to a depth of 2¾ inches. This will give a capacitance of about 35 pf.

Now hold the assembly over a low flame until the poly bubbles out of the holes in the aluminum tubing. When the poly has cooled, this will lock the pieces together so the capacitor won't fall apart.

The coils have 5 turns of No. 8 or No. 10 copper wire, 1½ inches inside diameter. The coil is fastened across the capacitor by means of clamps.

In this case the length of the coil was adjusted so that the trap resonated at 28.8 Mc. using a grid-dip meter for checking the frequency.

It was necessary to shorten the original 10-meter elements when the traps were added. The length of the driven element between outer ends of the trap coils was 14 ft. 2 inches, and the over-all length, including the 21-Mc. extensions (of ½-inch aluminum tubing) was 19 ft. 6 inches. Similar lengths for the reflector were 14 ft. 10 inches and 20 ft. 10 inches, while those for the director were 13 ft. 2 inches and 18 ft. 10 inches.

Feeding the Multiband Trap Yagi

The impedance at the center of the driven element will not usually be constant from band to band since the spacing between elements is not constant in terms of wavelengths. It can, however, be held to near enough the same figure to hold the s.w.r. within tolerable limits around the design-center frequency in each band, provided the element tuning is adjusted for the best compromise between reasonably-uniform feed-point impedance, gain, and front-to-back ratio. One method is to adjust for an impedance value

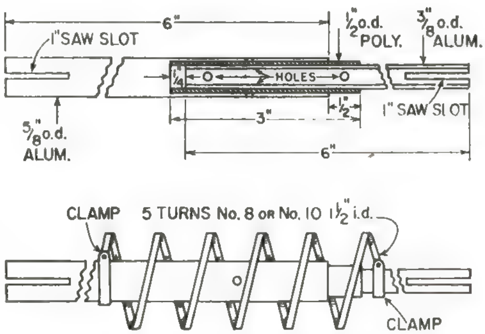


Fig. 12-17—Typical trap construction for Yagis with tubing elements (WØANY). Dimensions can be varied to suit the size of beam-element tubing. Increasing the diameter or length of the capacitor will increase the capacitance; increasing the radial spacing between the coaxial tubes will decrease the capacitance. This circuit was designed for modifying an existing 28-Mc. 3-element beam for use on 21 and 28 Mc. See text for antenna-element dimensions. Traps for 21-14 Mc. will require larger L and C values.

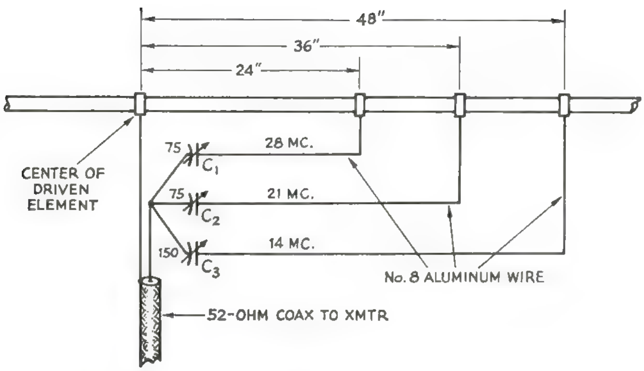


Fig. 12-18—Three-band matching system for trap Yagis. Use of other materials (e.g., tubing) for the gamma section probably will affect the dimensions shown; actual placement of gamma taps must be determined experimentally. Ordinary variable capacitors can be used if mounted in a weatherproof container; both the rotor and stator must be insulated from a metal container and an insulating shaft extension used for adjustment. Plate spacing should be 0.03 inch for power of 100 watts or so, increasing to at least 0.07 inch for a kilowatt. Spacing between the gamma wires or rods and the antenna element should be of the order of a few inches.

that will give a reasonable match to, say, 52-ohm cable, in which case the beam can be fed by opening the center of the driven element and connecting the line across the gap.

Another method that has been used successfully is shown in Fig. 12-18. It employs a separate gamma matching section for each band. The driven element therefore can be continuous and its center can be connected electrically to a metal boom. The dimensions shown were found to be satisfactory on a commercially-built three-band trap beam, and would not necessarily be exact for every trap system. The interlocking between adjustments for matching on the various bands was not found to be serious. The simplest method of adjustment is to match on the lowest frequency first, using an s.w.r. bridge, and work progressively to the highest band. Adjustment procedure in each case is as described previously for the single gamma, and after going through the complete process once the adjustments should be rechecked on each band.

MULTIPLE-YAGI MOUNTING

The beams described in the preceding section of this chapter illustrate typical constructional methods which are applicable not only to the single-band antennas shown but also to various multiband systems that have been used from time to time. One obvious way to secure directivity and gain on two or three bands is to build a separate antenna for each band, stacking all of them on a single vertical support and rotating the group as a unit. Equally obviously, this introduces mechanical problems of considerable magnitude, as well as being a relatively expensive approach. However, it has been done; furthermore, it is probably the most effective method of achieving multiband operation since it requires no sacrifice of electrical performance.

When individual Yagi antennas are stacked in this fashion, experience has shown that the minimum separation between antennas, to avoid interaction and possibly deterioration of performance of one or more of the beams, is about 0.1 wavelength at the lower of the two frequencies. That is, a 14-Mc. beam should be separated by at least six feet from a higher-frequency antenna, a 21-Mc. beam should be mounted four feet or more from a 28-Mc. antenna, and so on. In every case the antennas should be mounted so their elements are parallel, rather than arranged so that the elements of one are at some angle with respect to the elements of the next. The highest-frequency antenna should be at the top of the stack, the lowest-frequency one at the bottom.

Where a separate antenna is used for each band the design follows the same principles as in the case of single-band antennas, assuming that separations of at least the values just mentioned are maintained. In general, it is also necessary to use a separate matching system and a separate transmission line for each antenna.

Interlaced Elements

Another method that has been used is that of "interlacing" the elements for antennas designed for different bands by mounting them all in the same plane on the same boom. This is possible because the element lengths and spacings for consecutive amateur bands such as 21 and 28 Mc. are sufficiently different to make the interaction between the elements operating on different bands relatively small. A three-element beam for 28 Mc., for example, can have a two-element beam for 21 Mc. mounted on the same boom, using the space between the 28-Mc. elements for this purpose.

The antennas should be fed separately. In fact, except for sharing a common boom, the antennas should be designed independently. It is advisable to provide as much space as possible between the elements of one antenna and those of the one for the next band.

In adjusting arrays of this type, tune up the lowest-frequency antenna first, since it is likely to react more on the high-frequency antennas than the reverse. It is worth while checking the tuning of previously-adjusted antennas after each individual beam has been tuned.

Construction of Quad Antennas

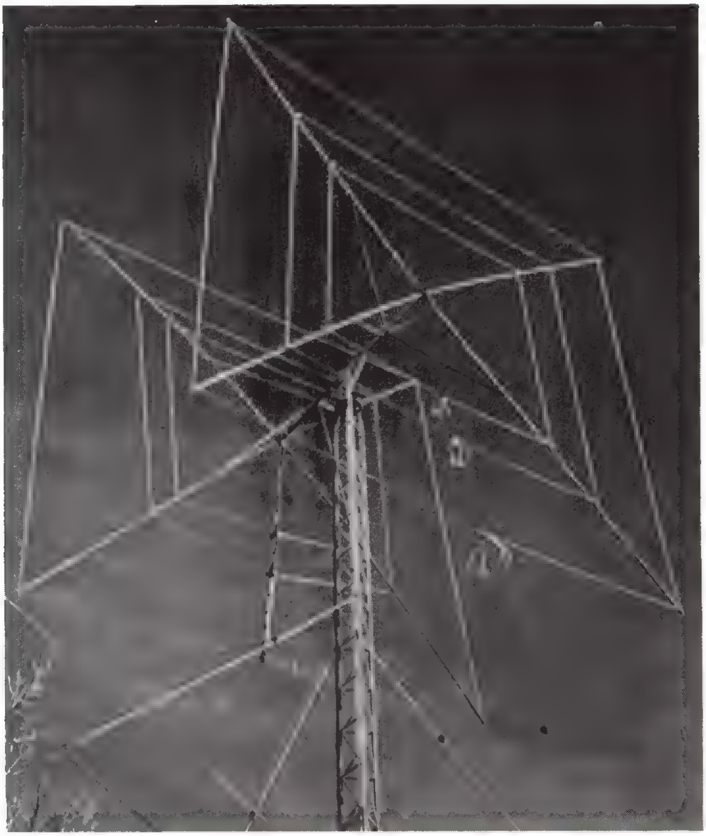


Fig. 12-19—Typical 2-element triband quad with compromise element spacing *(KØJTP)*

The construction of quad-type antennas involves the fabrication of a framework to support the wire loops, and a means of mounting the framework on a rotatable boom. The conventional framework consists of pair of spreaders (each usually made in two sections, or arms) joined together at the center to form an X, the four extremities of which support the four corners of the quad loop, as shown in Fig. 12-19.

MATERIALS

The quad loops are usually made of No. 14 or No. 12 enameled copper wire. Soft-drawn wire is preferable in this application, since the wire is not under significant tension. The tendency for hard-drawn or copper-clad steel wire to spiral makes it difficult to keep the loops in shape, unless the wire is prestretched.

Light-weight insulators should be used in suspending the wire loops from the spreader arms to minimize the weight that the spreaders must support. Small "airplane" type ceramic insulators, or strips of Lucite are commonly used.

The boom considerations are the same as they are for the Yagi types discussed earlier.

SPREADERS

In the past, the spreaders were almost exclusively bamboo poles, this material being nonmetallic and of light weight. Bamboo is still used frequently where minimum cost is essential. However, bamboo must be specially treated if it is to have a reasonable life span under outdoor weather conditions, and it is sometimes not easily obtained in the desired lengths and diameters. More recently, several manufacturers have brought out fiber-glass spreaders designed especially for constructing quad antennas. These spreaders are considerably stronger and require no weatherproofing. Aluminum spreaders have been used in some home-built quads, and at least one manufacturer produces a quad with spreaders of this type. However, so far as is known, the effects, if any, of all-metal spreaders in the immediate field of the antenna have not been closely evaluated. Because of this, metallic spreaders are sometimes broken up into two or three short sections joined together with short lengths of insulating material, such as waterproofed wood dowel.

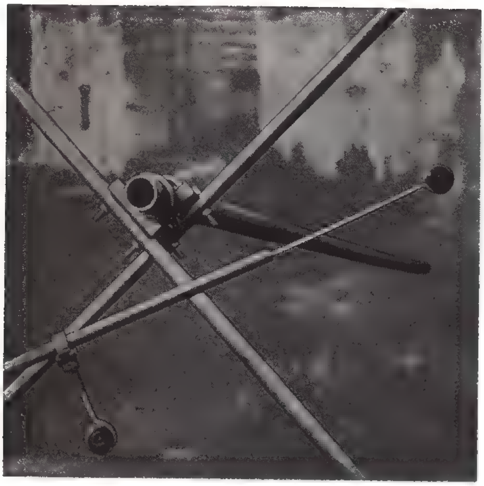


Fig. 12-20—Quad mount made from aluminum tubing and TV antenna fittings *(W8TUO)*. The aluminum struts are fastened to the boom with U bolts and saddles. This picture also shows the extension piece of one strut with TV standoff insulators that hold the quad wire.

Protecting Bamboo

The life of bamboo can be extended by applying two or three coats of marine varnish or boat resin, or by spraying with clear acrylic. Bamboo spreaders can be strengthened by covering them with a spiral winding of cotton rug-binders tape, masking tape, or fiber-glass tape before impregnating with varnish or spray. The diameter of bamboo spreader arms should be at least ½ inch at the tip, and preferably larger.

SPREADER MOUNTING

Table 12-I can be used to determine the total length of wire in a quad element. Dividing the result by 4 gives the length of each side of the loop. Multiplying this length by 1.41 gives the diagonal distance across the loop, and dividing this figure by 2 gives the approximate minimum length of each spreader arm.

Figs. 12-20 through 12-24 show various methods of mounting spreader arms on a boom. Fig. 12-20 is a method that can be used to mount metallic-tubing spreaders, or pipe sockets for metallic-tube, fiber-glass, or bamboo spreaders. This assembly can be strengthened, if need be, by leaving a 1/16 to ⅛-inch space between members where they cross, and running horizontal metal strip or bar from one spreader arm to the other, 6 to 12 inches both above and below the boom.

The method of Fig. 12-21 is quite similar except that ⅛-inch aluminum or steel angle stock, rather than tubing or pipe, is used for supporting the selected type of spreader. The two angle pieces should each be about 30 inches long, and should be of a cross section appropriate to permit clamping the bottom ends of the spreaders with stainless-steel hose clamps. A sketch of the aluminum plate on which the angle pieces are mounted is shown in Fig. 12-22. The boom rests in the large V-shaped notch. The notch should not be cut until the cross bars have been temporarily fastened on either side of the plate with a ¼-inch bolt at the center. Then the V can be scribed along the edges of the angle pieces. It is important that the edges of the V be filed flush with the angle pieces so that the angle faces will bear against the boom when the assembly is fastened to the boom. The "dog ears" at the corners of the plate help to keep the angle pieces from turning on the central bolt, since the hose clamps used at the bottoms of the spreaders also serve to clamp the angle pieces to the ears. The angle pieces are drilled for U bolts by which the assembly is fastened to the boom, as shown.

The spreader mounting shown in Fig. 12-24 makes use of a "Kee clamp." These clamps are essentially blocks of iron or steel, drilled and tapped with five holes, as shown, and are designed for use in assembling pipe scaffolding. All holes are usually of the same size. A size should be selected that will permit threading the central hole onto the end of a pipe boom (steel or aluminum), Standard pipe "nipples" are then used to reduce the peripheral holes to a size suitable for pipe stubs or sockets for the spreader arms. In Fig. 12-24, the Kee clamp is for 1½-inch pipe, and stubs are provided in the peripheral holes for mounting spreaders made of electrical conduit. The Kee clamps come in a variety of standard pipe sizes, and it is obvious that details can be varied to suit booms of other sizes, and other types of spreaders.

MULTIBAND QUADS

The same spreaders may be used to support elements for two or three bands, as shown in Fig. 12-19. In this instance, however, the element spacing in terms of wavelength cannot be the same for all bands, since the spacing is fixed in terms of feet. Therefore, the spacing must be a matter of compromise. A distance of 7 feet between the driven-element spreaders and the parasitic-element spreaders results in 0.1 wavelength spacing on 14 Mc., 0.15 wavelength spacing at 21 Mc., and 0.2 wavelength at 28 Mc.

TABLE 12-I

Quad Element Dimensions
Length in feet for Complete Loop

Driven element	–	$\frac{1005}{f_{Mc.}}$
Reflector	–	$\frac{1030}{f_{Mc.}}$
Director	–	$\frac{975}{f_{Mc.}}$

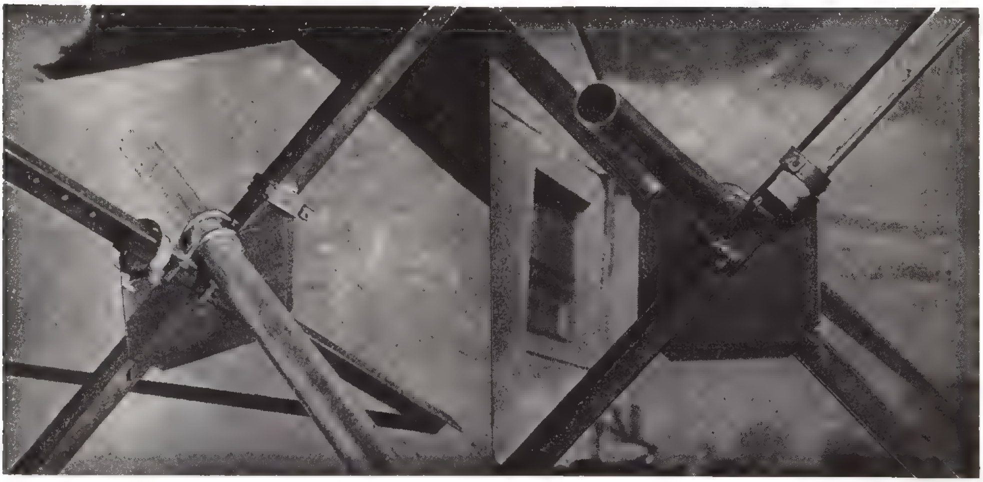


Fig. 12-21—Inside (left) and outside (right) views of a spreader support for quad antennas (*W3PRU*). The assembly is clamped to a tubular boom with U bolts. The spreaders (bamboo shown here) are clamped inside angle-stock crossbars with adjustable hose clamps, both near the center of the assembly, as shown, and at the outer ends of the angle pieces.

Fig. 12-25 shows a constructional design that permits constant spacing of elements for a two-element three-band quad. This design requires no boom. It can be seen that the spreaders for each set of elements are not all in the same plane, as they are in Fig. 12-19, but are angled outward from the center to form the outline of a pyramid on either side of the mast, with the apexes of the pyramids at the mast. The 20-meter elements are supported at the outer tips of the spreaders, while the higher-frequency elements are suspended at points closer to the mast. In the arrangement shown in Fig. 12-25, the spreaders are mounted in pipe sockets (or fastened to sections of angle iron) which are welded to a vertical pipe sleeve that mounts on the mast. The angles at which the sockets are mounted must be such that the separation at the ends of the spreaders will be equal to the length of one side of the quad loop for the lowest frequency, plus allowance for insulators and fastening hardware. A further requirement is that this spacing must occur at a horizontal distance from the mast that will result in the desired spacing between the lowest-frequency driven element and parasitic element. When this is done, there will be sufficient space between spreaders for higher-frequency elements at a similar element separation in terms of wavelength.

A simpler arrangement permitting experimental adjustment of the angles involved is shown in Fig. 12-26. Four pieces of pipe or tubing into which the spreaders will fit are bent as shown at A. These pieces are then clamped to a heavy plate by means of U bolts with serrated yokes, as shown at B, staggering the U bolts holding the tubing member mounted on the top of the plate with those holding the

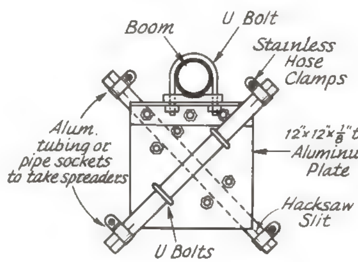


Fig. 12-23—Another type of quad-spreader mounting. Pipe sockets for the spreaders are fastened by means of U bolts to a plate suspended from the boom. The plate is attached to the boom making use of an angle bracket, and an additional U bolt (*W6JIC*).

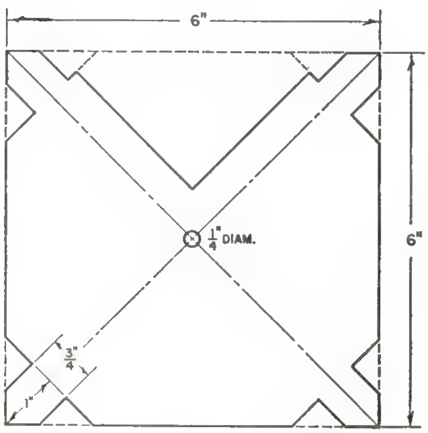


Fig. 12-22—Dimensions of the aluminum plate that supports the angle pieces. The large V notch should not be cut until the angle pieces are in place.

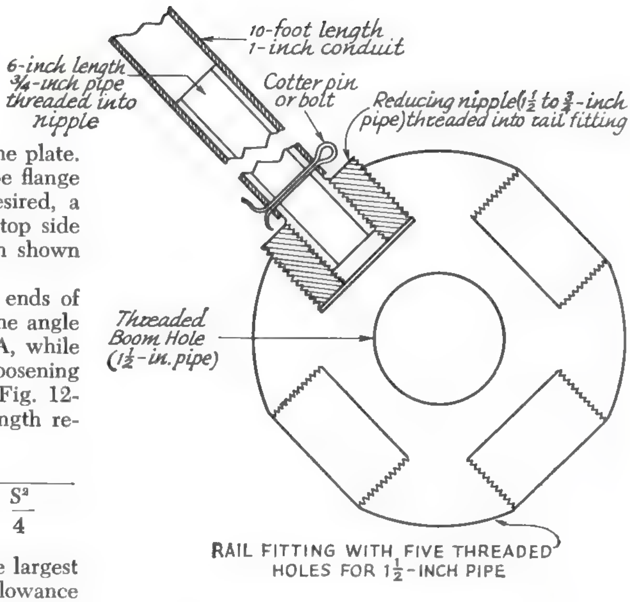


Fig. 12-24—This quad-spreader support makes use of "Kee clamps" of the type used in assembling pipe scaffolding. (KØJTP)

member fastened to the under side of the plate. The plate, in turn, is mounted on a pipe flange fastened to the top of the mast. If desired, a second flange may be mounted on the top side of the plate to carry the mast extension shown in Fig. 12-25.

The horizontal distance between the ends of the spreaders is adjusted by adjusting the angle of the bends at Angle A in Fig. 12-26A, while the vertical separation is adjusted by loosening the U bolts and changing Angle B in Fig. 12-26B. The approximate spreader arm length required may be determined from

$$\textit{Spreader arm length (ft.)} = \sqrt{\frac{E^2}{2} + \frac{S^2}{4}}$$

where E is the length of one side of the largest element to be mounted (in feet, plus allowance for insulators), and S is the desired distance between the largest driven element and its parasitic element (in feet). Spreaders about 15 feet long are suitable for a separation of 20 feet at the ends of the spreaders, at a horizontal distance of 5 feet from the mast. The total separation between driven and parasitic elements will then be about 10 feet, or slightly over 0.15 wavelength at 14 Mc.

If the 28- and 21-Mc. elements are then placed 2½ feet, and 3¾ feet, respectively on either side of the mast (horizontal measurement), these elements will also be spaced about 0.15 wavelength, and the distance between spreaders at this point will be appropriate for the lengths of wire required for the quad loops for these bands.

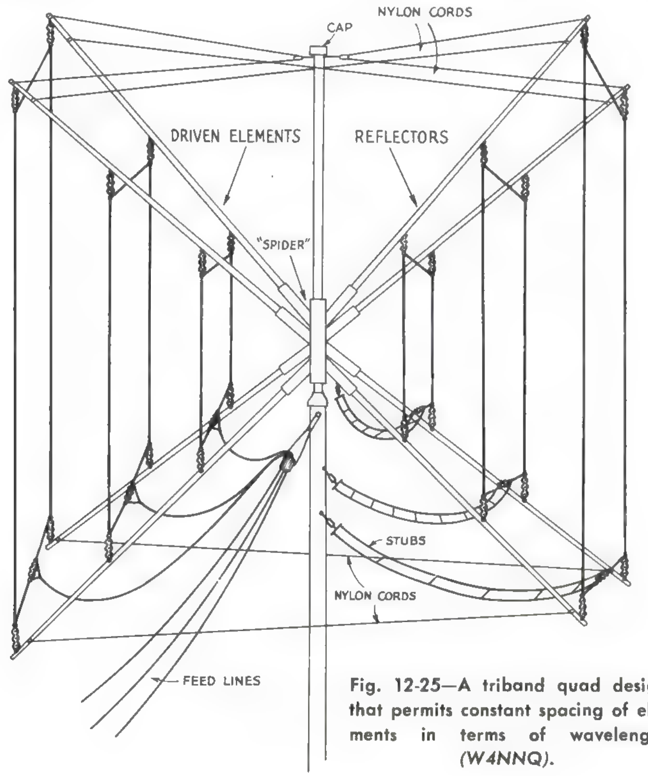


Fig. 12-25—A triband quad design that permits constant spacing of elements in terms of wavelength (W4NNQ).

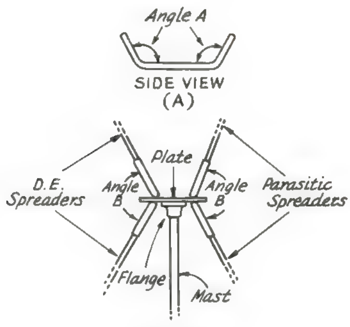


Fig. 12-26—Sketch showing the essentials of another type of spreader mounting for a multiband quad with constant-wavelength element spacing. See text for details.

FEEDING QUAD ANTENNAS

The feed-point impedance of a 2-element quad will vary somewhat with element spacing, but will usually be in the vicinity of 100 ohms, so the mismatch to 75-ohm coax line will not be prohibitive. However, the impedance can be matched more closely by means of a gamma matching section or, more simply, by using a quarter-wave transformer. For a feed-line impedance of 50 ohms (RG-8/U), this transformer may consist of a section of RG-11/U cut to a length of $162.4/f_{MC}$ feet. This section is inserted between the feed point of the antenna and the RG-8/U line.

In the case of multiband quads, it is preferable to use a separate line to each driven element. However, single-line feed, with the feed points of all driven elements connected in parallel, may be used where economy is essential. In some instances, single-line feed has been used with a separate gamma match for each driven element, the feed points of the gammas being connected in parallel.

THE "SWISS" QUAD

The name "Swiss Quad" has been given to the antenna configuration shown in Fig. 12-28 by its designer HB9CV. It differs from the conventional quad electrically in that both elements are driven—with a phase difference of 180 degrees. Construction is simplified by making the horizontal members of aluminum tubing sufficiently rigid to support the weight of the vertical members, which are made of wire, thereby eliminating the customary diagonal spreaders. Additionally, the horizontal members are bent in such a way as to provide the desired element spacing without the need for a boom. So far as is known, no close comparison between the performances of this antenna and the conventional quad have been made, but good results have been reported from many sources.

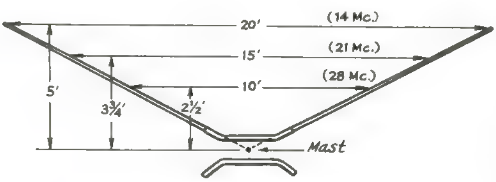


Fig. 12-27—Sketch showing element mounting dimensions that result from the example given in the text. This may be considered to be a view looking down on one pair (top) of spreaders from above, although the spreader lengths are not to scale, since they slant upward toward the viewer, as well as slanting outward as shown.

The horizontal members must be insulated from each other, and from the mast, if it is of metal, as shown in Fig. 12-29. High-grade insulation is not required. The center points of each

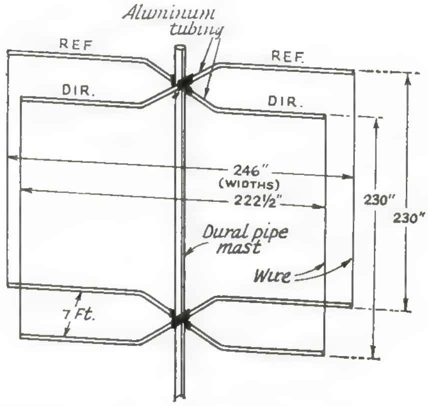


Fig. 12-28—The "Swiss Quad" antenna needs no spreaders or boom. Dimensions shown here are for an operating frequency of 14.25 Mc. See Table 12-II for other frequencies.

pair of horizontal members should come as close together as constructionally possible. These points should then be connected together, and a connection to the mast made with a common lead. All leads should be as short as possible.

The tips of the horizontal members should be flattened and drilled for bolts for attaching the vertical wires, which may be of copper, although aluminum wire should be equally satisfactory.

The dimensions shown in Fig. 12-28 are for 14.25 Mc. Table 12-II show suggested dimensions for some other frequencies.

The horizontal members are made as shown in Fig. 12-30. A single length of aluminum tubing (⅞ inch o.d., 18-gauge wall is suitable) is used for the center section. Smaller tubing is used for the end sections. Since one of the extensions is in the "director" element, while the other is in the "reflector" element, the extensions are not of equal length.

Fig. 12-31 shows a method of feeding the antenna. Before connecting the gamma section,

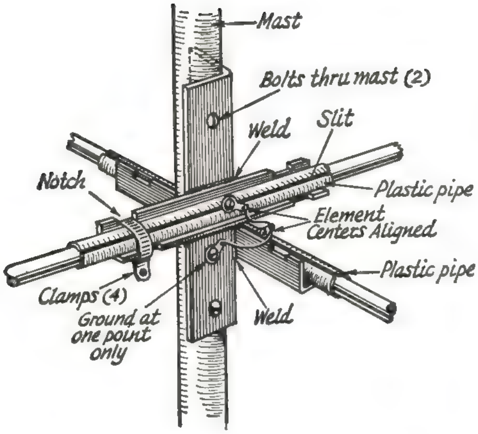


Fig. 12-29—The elements are insulated from the angle-stock mounting bracket by sections of plastic pipe that have been slit. The connections shown, made with sheetmetal screws through the slit opening, should be made at the exact center points of the elements. These points are connected to the mast with a single short common lead. The braket might also be made by assembling the angle pieces with bolts, instead of welding.

TABLE 12-II

Typical Swiss Quad Dimensions

Freq. Kc.	*Height*	*Reflector Width*	*Director Width*	*Spacing (0.1λ)*
28,500	116	121.5	110	41.3
21,200	156	164	148	55.5
14,150	234	246	222	83.5
7,050	470	493	443	168

All dimensions are in inches, and are based on a design reflector perimeter of 1.148 wavelength, and director perimeter of 1.092 wavelength, suggested as optimum by HB9CV. Width is the overall length of the horizontal members, as indicated in Fig. 12-28. Height should be adjusted for antenna resonance at the desired center operating frequency.

a small loop of wire should be connected across the feedpoint. The lengths of the vertical wires should then be adjusted equally (raising or lowering the lower set of horizontal members to compensate) until a grid-dip oscillator coupled to the loop indicates resonance at the desired frequency. Then the gamma section should be mounted, and its length adjusted equally each side of the crossover point for minimum s.w.r. A gamma spacing of 1/200 wavelength is suggested by HB9CV and an initial setting of the taps at about half way between the bends and the ends of the horizontal members, but keeping the taps at equal distances in respect to the crossover point. The gamma section may be made of wire suspended from the aluminum tubing by means of insulators of Lucite strip fitted with clamps for the tubing.

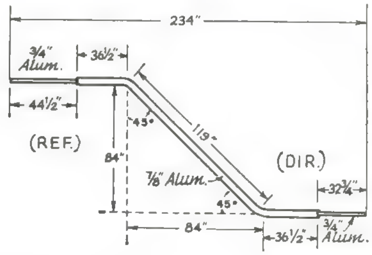


Fig. 12-30. The bends in the center section of the element should be made with a conduit bender. Dimensions shown here are for 14.25 Mc. Since the halves of the section are in different elements, the extensions are not of equal length.

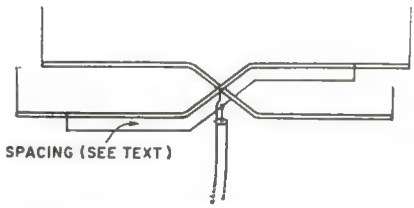


Fig. 12-31—Matching system for coaxial line. The outside conductor of the line is connected to the grounding lug on the mast (Fig. 12-29). The center conductor is connected to the center of the gamma.

Supports for Rotatable Antennas

While masts of the type discussed in Chapter 11 as supports for wire antennas are sometimes used for mounting rotatable antennas, they have obvious disadvantages in the latter application. They cannot be climbed, nor can the antenna usually be lowered readily when service becomes necessary. A more serious drawback is that a mast has limited ability to withstand the torsion that develops with wind pressure on the antenna, and when rotation of the antenna is stopped and started. Even if the mast is sufficiently sturdy to withstand these stresses, there still remains a tendency for the antenna to oscillate back and forth after rotation has stopped, and also in the presence of wind. This makes it difficult to establish and maintain a desired directional heading.

Tilt-Over Mast

A 60-ft. tilt-over mast that is capable of supporting antennas up to triband Yagi size is shown in Figs 12-32 through 12-35. It requires no guying, and thus can be erected in small space. Oscillation is minimized by the use of sections

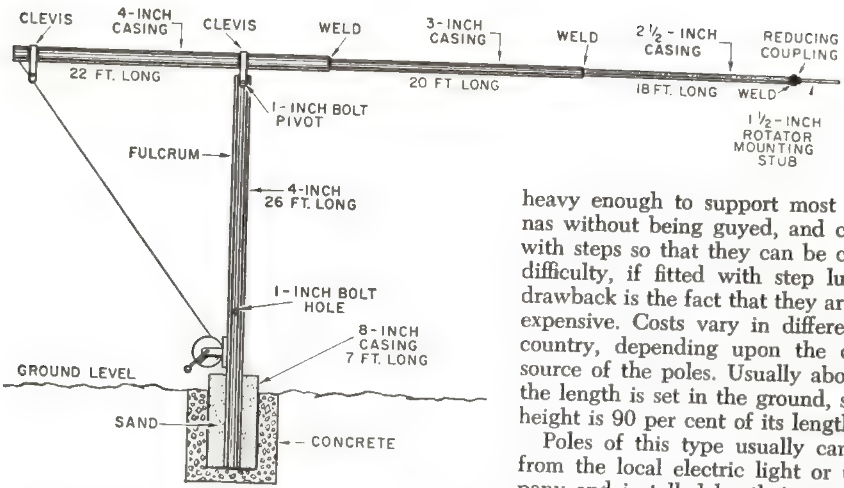


Fig. 12-33—Sketch showing the over-all design of the steel tilt-over mast. Sections are telescoped about 4 inches before welding.

of much larger cross section than is customary in masts used for supporting wire antennas.

The mast shown is made of steel well casing obtained from a junk yard where welding facilities were also available. Referring to Figs. 12-33 and 12-34, the foundation consists of a 7-ft. length of 8-inch well casing set in concrete to a depth of 6 ft. below ground level. The fulcrum is a 26-ft. length of 4-inch casing held central inside the base section while the space between the two is filled with sand.

The mast proper consists of a 22-ft. length of 4-inch casing, a 20-ft. length of 3-inch casing, and an 18-ft. length of 2½-inch casing. The top end of the 2½-inch section terminates in a reducing coupling and a short length of 1½-inch pipe on which the rotator is mounted. The antenna is supported on a 6-ft. stub of 1½-inch pipe mounted in the rotator socket. All joints between sections, including the one at the reducing coupling, are welded.

At a point 15½ ft. from the bottom end of the bottom mast section, a clevis made of ½ × 2-inch steel strap is centered and welded. Holes to clear a 1-inch bolt are drilled in the clevis and at a point 2 inches down from the top end of the fulcrum, as shown in Fig. 12-35. A similar clevis is welded a few inches from the lower end of the bottom section of the mast, and matching holes are drilled in the clevis and fulcrum. The complete assembly should be given a coat of aluminum paint.

The mast is raised and lowered by means of a boat winch mounted near the base of the fulcrum. One person can operate the winch quite easily.

TELEPHONE POLES

Probably the most satisfactory type of mast is the kind used by utility companies to carry power and telephone wires. These poles are heavy enough to support most amateur antennas without being guyed, and can be provided with steps so that they can be climbed without difficulty, if fitted with step lugs. Their chief drawback is the fact that they are comparatively expensive. Costs vary in different parts of the country, depending upon the distance to the source of the poles. Usually about one-tenth of the length is set in the ground, so that the pole height is 90 per cent of its length.

Poles of this type usually can be purchased from the local electric light or telephone company and installed by their crews. In some localities the companies let out this work to local contractors, in which case the contractor can be approached directly. Use caution in picking up "bargains" in the pole line; a pole with an unsound center is not a good investment nor is it a safe thing to climb. Fig. 12-36 shows a method of mounting a rotator on a pole of this type.

TOWERS

By far the largest percentage of amateur rotatable antennas are supported on manufactured towers. These towers come in a wide

Fig. 12-32—A 60-ft. tilt-over mast made of steel well casing. The mast will support a triband beam without guying. *(KØONM)*.

Fig. 12-34—Close-up view of the base, showing the sand packing between the base mounting and the fulcrum, the hoisting winch, and the bottom clevis which secures the lower end of the mast to the bottom of the fulcrum.

variety of styles, heights, and wind-load ratings. In addition to the rigid types that are assembled in prefabricated sections, usually 10 ft. per section, there are types that can be raised or lowered, by the operation of a winch at the base, to bring the antenna down to a level that can be reached with a ladder, making it unnecessary to climb to the top of the tower to mount or service the antenna. This feature also makes it possible to shorten the tower for added safety in periods of abnormally high winds. Most types are lowered by telescoping sections, although there is at least one nontelescoping type that permits lowering the antenna by virtue of a hinge at the approximate center point of the tower. With some telescoping types it is also possible to hinge the base section over to ground level after the upper sections have been telescoped. Some tower manufacturers also offer an electric-driven winch to replace the usual manually-operated winch.

Manufactured towers are shipped in knockdown form, complete with all hardware necessary and detailed instructions for assembly. The instructions include specifications for the concrete base and for guying, if guying is necessary. Most telescoping towers are designed to be free-standing, without guy wires, if the wind load is restricted to a specified maximum. However, the permissible antenna area can be at least tripled if guy wires are used. Guy wire and all fittings necessary for guying can usually be obtained from the tower manufacturer in kit form, making it unnecessary to shop around for this material.

Tower Wind Load

The wind load that an antenna imposes on a tower is expressed in pounds and, at any instant, is the product obtained by multiplying the surface area of the antenna in square feet by the wind pressure in pounds per square foot. To comply with accepted standards, the antenna wind-load rating of a tower should be, either directly or indirectly, in terms of the permissible antenna area for the maximum wind pressure to be expected in the locality where the tower is to be used. For most of continental U.S.A. the accepted standard for maximum wind pressure is 30 lbs./sq. ft. However, for most regions along the Gulf Coast and Atlantic Seaboard, this figure rises to 40 lbs./sq. ft. In certain coastal sections of Texas, Louisiana, Florida, and North Carolina, where winds of hurricane force may be expected, the figure rises again to 50 lbs./sq. ft. (If there is any doubt about the rating to be used in a certain locality, consult local building authorities.)

The foregoing figures are for flat surfaces. For a given wind velocity, the pressure developed against cylindrical surfaces (this applies to most rotatable antennas, of course), is approximately $2/3$ of that developed against flat surfaces. Therefore, when cylindrical antenna members are used, the pressures mentioned above can be reduced to 20, 27, and 33 lbs./sq. ft., respectively. This is equivalent to an increase of 50% in permissible antenna area.

The antenna area specified for a tower in a 30-lb./sq. ft. region must be reduced in inverse proportion for the higher-pressure regions. For example, in a 50-lb./sq. ft. region, the antenna surface area should be reduced to 30/50 or 3/5, of the flat area permissible in the 30-lb./sq. ft.

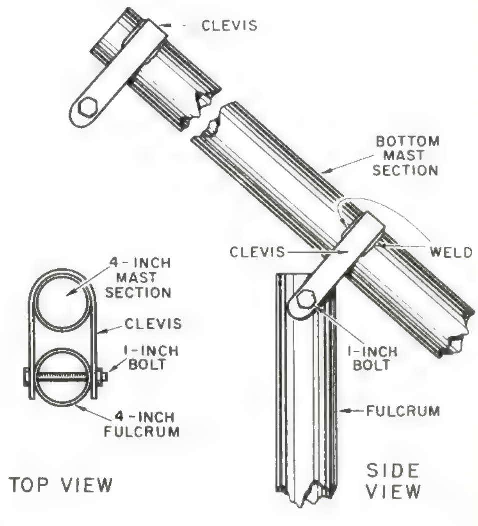


Fig. 12-35—Detail sketch showing how the top clevis is used to provide tilting action.

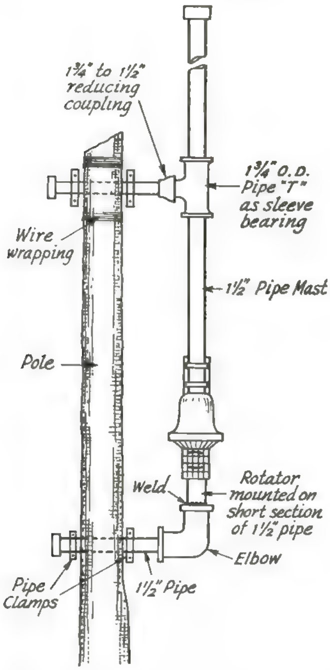


Fig. 12-36—A simple method of mounting a rotator on a telephone pole. All fittings are readily obtainable as a plumbing shop.

region. The area permissible with cylindrical members will also be ⅗ of the cylindrical area permissible in a 20-lb./sq. ft. region.

Some tower manufacturers rate their towers directly in terms of permissible antenna area at stated wind pressures. Others use a rating in terms of total wind-load pressure in pounds (this should not be confused with the figure for antenna weight), while still others refer to the permissible antenna area at some maximum wind velocity in miles per hour.

Total wind load in pounds is the product obtained by multiplying the antenna area in square feet by the allowable wind pressure in pounds per square foot. Therefore, to obtain the permissible antenna area, it is necessary to divide the total wind-load figure by the maximum allowable wind pressure in pounds per square foot. As an example, for a tower with a total wind-load rating of 200 lbs., the permissible antenna area for a maximum pressure of 20 lbs./sq. ft. (cylindrical members in a 30-lb. /sq. ft. area) would be 200/20=10 sq. ft.

Wind velocity is related to pressure approximately by:

Pressure (lbs./sq. ft.) $= 0.0025V^2$

for cylindrical surfaces, where V is the velocity of wind in miles per hour. Thus, a pressure of 20 lbs. per square foot on cylindrical surface results from a wind velocity of about 90 m.p.h.

If the tower literature does not contain the preceding information, it should be requested from the manufacturer before purchasing a tower, since this data is necessary in determining whether or not a particular tower is adequate to handle the desired antenna load. A reputable tower manufacturer will also supply any other design specifications that a local building inspector may request.

Antenna Wind-Load Area

The wind-load area of a Yagi antenna can be estimated by multiplying the length of the longest element by its average diameter (both measurements in feet), and multiplying again by the number of elements. The area of traps, if any, should be calculated separately, and added in.

The wind-load area of a quad can be similarly estimated by multiplying the length of one spreader arm by its average diameter (both in feet), multiplying by 4 (for four arms), and multiplying again by the number of elements. To be safe, about 20% should be added to take care of the area of the wire, and the mast that is generally used with a quad, and the fact that the length of the mast is practically equivalent to an increase in tower height, so far as the resulting stresses on the tower are concerned.

Rotating the Antenna

The majority of rotatable installations use motor drive as the means for turning the antenna, but rotation by hand is not at all impracticable if the mast holding the antenna is rigged in such fashion as to be capable of being rotated. If hand rotation is to be a reasonably convenient operation it is necessary that the mast be within easy reach of the operator, although a mechanical drive using belts or chains can be devised if the mast has to be several feet away from the operating position—on the outside wall of the house, for instance.

Another possibility, when the operating position is on a floor directly underneath the roof of the house, is to use a pipe mast projecting through the roof from the station, using a sleeve bearing at the roof. Care must be used in weatherproofing the roof in such a case, of course.

MOTOR DRIVE

"Heavy-duty" TV-antenna rotators will handle any amateur antenna of a size comparable with the larger arrays used for TV reception. These rotators are quite suitable for many v.h.f. beams, therefore, and for the smaller varieties of beams designed for the lower frequencies, such as a three-element 28-Mc. beam or even a "trap" beam for 14-28 Mc.

Rotators of this type readily can support the weight of an amateur antenna, but this is a relatively minor consideration compared with the necessity for withstanding the stresses occa-

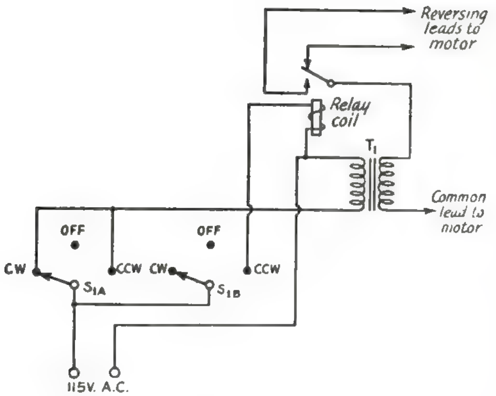


Fig. 12-37—A method for using only three wires to control a reversible propeller-pitch motor used as rotary-beam drive unit. The d.p.d.t. switch has a neutral, or off, position. The relay is a 115-volt a.c. s.p.d.t. type.

sioned by the forces exerted on the beam by wind. As the over-all effect of wind depends on the exposed area of the antenna, caution should be used in selecting a rotator for an antenna of large dimensions, especially one using a large-diameter boom and elements. A full-size 14-Mc. beam, for example, would require a rotator designed to handle much larger wind forces than could be met safely with TV rotators, even though the latter could support the weight and would be capable of turning the antenna on a calm day.

Rotators designed either for TV or amateur work usually include some method of braking, to hold the antenna in position once it has been turned to point in the desired direction, and also have built-in provision for preventing rotation beyond 360 degrees. Bringing off the feeder from the antenna thus presents no special problems, especially with a coax line, since it is only necessary to provide enough slack in the line to allow the antenna to turn through one complete circle. The usual speed of rotation is between ¾ and 2 r.p.m. in commercially-available rotators.

Home-Made Drives

In assembling a motor-drive system from available parts, the principal difficulty is likely to be the acquisition of a suitable gear train. The ultimate turning speed of the antenna should be of the order of 1 r.p.m. The required gear reduction depends on the motor speed which, in the case of ordinary a.c. motors of the type used in household appliances, is about 1750 r.p.m. Gear trains can be assembled from gears purchased new for the purpose, but it is also possible to find suitable ones in discarded appliances —for example, the gears used in washing machines.

The builder usually has to exercise some ingenuity in devising a mechanical system that permits adapting materials that may be useful in a rotator. Also, it is necessary that a weatherproof housing be constructed, and provision must be made for reversing the direction of rotation, either by electrical or mechanical means. The complete installation should include a method for preventing the antenna from going beyond the limits of one complete turn, a function that, for example, can be performed by limit switches incorporated in the rotating mechanism.

Not a great deal of power is needed: a motor capable of delivering ⅛ horsepower is ample. The reduction gears will act as a brake for the system so that it does not "coast" too far past the position at which the power was shut off.

If the motor is a low-voltage high-current unit, as are the popular "propeller-pitch" motors to be found in war surplus, the step-down transformer should be mounted near the motor (if a long run of power line is required), to avoid the use of unnecessarily heavy lines. Control switching can then be done with a heavy-duty relay at the transformer, as shown in Fig. 12-37.

A beam is generally positioned by rotating it to the bearing that gives the loudest received signal. To do this, the rotating motor must radiate little or no electrical noise, of course, and this requires some type of filtering at the brushes of a commutator-type motor. Shielding the leads to the motor and grounding its frame are also helpful in some cases. The noise generated by the propeller-pitch motors can be eliminated by using six small capacitors, as shown in Fig. 12-38. First remove the thin aluminum motor cover and then take out the motor itself. After cleaning the six copper-surfaced brush holders and drilling and tapping three holes for 6-32 screws, solder a capacitor from each brush holder to a ground lug at the tapped holes. Mica or ceramic capacitors can be used, in the range 0.002 to 0.01 μf. The disk ceramic type is good for the job because of its small size.

ROTATOR MOUNTING

The best method of mounting a rotator on a tower is shown in Fig. 12-39A. The thrust bear-

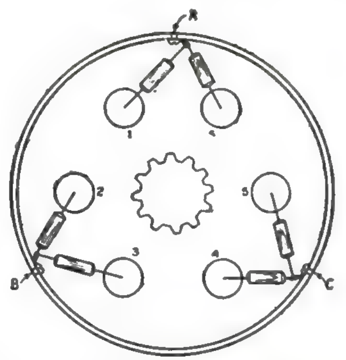


Fig. 12-38—Noise created by propeller-pitch beam rotators can be eliminated by bypassing the brush holders to the case of the motor as shown. Points A, B and C are grounds made by drilling and tapping the rim of the motor case for 6-32 screws.

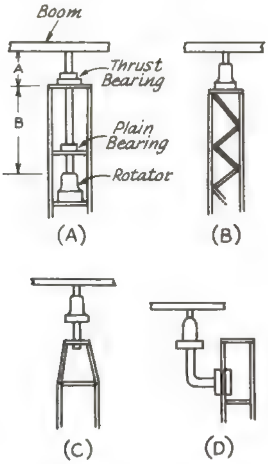


Fig. 12-39—Methods of mounting a rotator at the top of a tower.

ing relieves the rotator of the weight of the antenna and mast, while the plain bearing removes the side pressure from the rotator bearings. This arrangement is particularly desirable if the mast is to extend more than a foot or two above the top of the tower. A less-effective, but sometimes adequate, method is to omit the lower bearing, and use a plain bearing at the top. In this case, the rotator bearings carry the full weight of the antenna and mast. The side thrust on the rotator bearings depends on the ratio of length A in Fig. 12-39A to length B, the pressure decreasing as this ratio is made larger.

The top section of the tower may not have sufficient cross section to accommodate the desired size of rotator. In this case, the rotator can be mounted on the top plate of the tower, as shown in Fig. 12-39B, and the mast, which should be as short as possible, coupled directly to the rotator. Some tower manufacturers provide a pipe stub in the top of the tower, or offer an offset stub that may be clamped to one of the tower legs, as shown in Figs. 12-39C and D. In all three latter methods, the antenna weight and side stresses are imposed directly on the rotator bearings. The methods shown in C and D also involve additional points where slippage may occur.

With rigid towers, the rotator is sometimes placed at ground level, the antenna being mounted on a shaft that runs the full length of the tower. While this provides easy access to the rotator for servicing, it suffers from the same effects of torsion mentioned earlier in reference to mast supports.

INDICATING BEAM DIRECTION

Most users of beam antennas like to know which way the beam is pointing without having to resort to telescopes, periscopes and other optical devices. If the beam is manually operated it is a simple matter to make a string- or cable-driven indicator along the lines of the schemes shown in Fig. 12-40.

TV rotators usually are sold with a control and indicator unit as part of the complete package. Many of these use indicators that are essentially voltmeters calibrated in terms of direction, the voltage applied to the meter being taken from a potentiometer ganged with the antenna drive shaft. Fig. 12-41 is a basic circuit of this type. Ordinary potentiometers do not have full 360-degree rotation and therefore are not entirely suitable (the ones used in commercial rotators are special items) although they can be used if the rotating system has limit switches or other means for preventing rotation beyond the angular distance that the potentiometer can handle.

In Fig. 12-41, if T_1 is a 6.3-volt filament transformer, M_1 can be a 0-5 volt a.c. voltmeter. R_2 should be adjusted so that the voltmeter reads full scale when the potentiometer arm is at the maximum-voltage end (fully clockwise in Fig. 12-41). To allow for adjustment, the maximum value of R_2 should be at least ⅓ the resistance of R_1, and the total of the two resistors should be such as not to overload the transformer. If the current taken by M_1 is small compared with the current through R_1 and R_2 the indications will be directly proportional to the angular travel of the potentiometer arm.

A deluxe system of beam-direction indication uses "synchro" (Selsyn) generators and motors, as shown in Fig. 12-42. These a.c.-operated devices are available for 60-cycle 115-volt operation and for 400-cycle 115-volt operation. The 400-cycle units are less expensive (in war surplus) as a rule, and can be used with a 60-cycle supply at reduced voltage.

However, there is a reduction in accuracy (as much as 20 degrees in some instances, and the accuracy is likely to be sluggish or jerky. Fig. 12-43 shows the circuit of simple supply that can be built to convert 60-cycle line supply

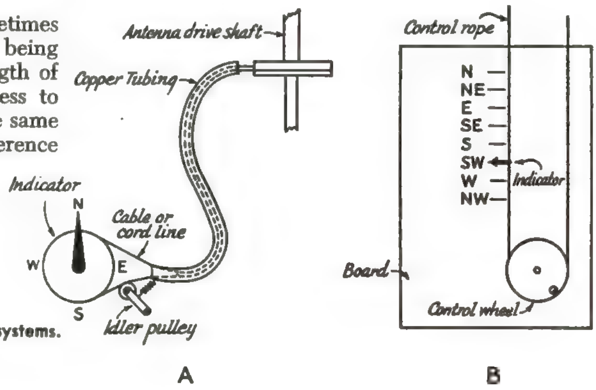


Fig. 12-40—Mechanical direction-indicator systems.

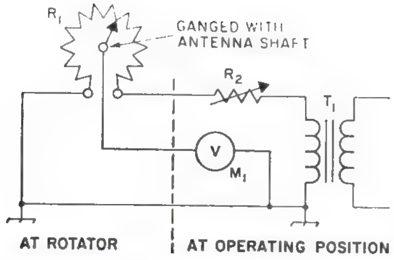


Fig. 12-41—Voltmeter-type indicator circuit.

M_1—A.c. voltmeter (0–5 range if T_1 is a 6.3-volt filament transformer).

T_1—Step-down transformer; 6.3-volt filament transformer suitable.

R_1—Wire-wound potentiometer, 20 ohms or more; see text.

R_2—Wire-wound slider type or rheostat, ⅓ to ½ resistance of R_1.

to 400 cycles at a voltage suitable for operating surplus selsyns.

The 2N176 transistor is shown in the diagram, but any similar transistor should prove satisfactory. The value of the resistor R_1 is not critical. A value of 390 ohms is suggested as giving the highest output voltage. The critical frequency-determining components are the output transformer T_1, and the capacitors C_1 and C_2. If a transformer other than that suggested is used, the capacitors must be adjusted to compensate for any change in transformer impedance. Best results are obtained by fixing C_1 at about 10 μf., and then varying the value of C_2 until an output frequency of 400 cycles is obtained. Reducing the value of C_2 increases the output frequency, while an increase in the value of C_2 lowers the output frequency. Tests with several different output transformers resulted in using values of C_2 ranging from 4 to 20 μf., depending upon the particular transformer.

While voltages across C_1 and C_2 are quite low, capacitors of very small physical size tend to heat, resulting in frequency instability and eventual breakdown. Capacitors in the range of 50 to 150 w.v.d.c. should be satisfactory.

Frequency Adjustment

Checking the output frequency of the unit can be accomplished by either of two methods. First, and most desirable, is by the use of an oscilloscope and a calibrated audio oscillator. In this method of adjustment, the output of the power supply is fed to either the vertical or the horizontal plates, while the output of the audio oscillator at 400 cycles is fed into the other scope plates. When the value of C_2 is correct, the usual circular pattern will appear on the scope. The figure will probably be a flat-sided circle more nearly resembling the letter D than the letter O.

The second method is by the use of a speaker or headphones across the output of the supply (be sure and connect a capacitor of about 0.2 μf. in series with the speaker-transformer primary or phones) and comparing the output frequency by ear with that from a tuning harp or fork or, even better, by beating against the 440-cycle transmission from WWV.

In the final adjustment of the value of C_2, it is desirable that the 400-cycle power supply be connected either to the Selsyns it is intended to operate, or to a similar pair.

The 10 to 14 volts at 60 cycles required to operate the power supply can be obtained from a tap or winding on the transformer operating the antenna rotator, or from a small low-current transformer such as a bell-ringing transformer.

There is no difference in operation in the two systems in Fig. 12-42 except that Circuit B requires one less wire. In Circuit A or B the relative directions of rotation can be reversed by interchanging the connections of two of the stator (S) connections at one unit.

The generator synchro should be coupled to the antenna through a 1-to-1 ratio unless a duplicate gear system is used at the motor (indicator) end. The coupling can be through gears, pulleys and belts, or friction drive.

SELSYN-DRIVEN DIRECTIONAL INDICATORS

Many different ways have been devised for indicating the beam direction at the operating position when using Selsyns, ranging from simple pointers moving over a compass chart to back-lighted world maps and rotating globes. An example of the former is shown in Fig. 12-44. This indicator has a great-circle map sandwiched in between two ⅛-inch pieces of Plexiglas. The map is 12 inches in diameter,

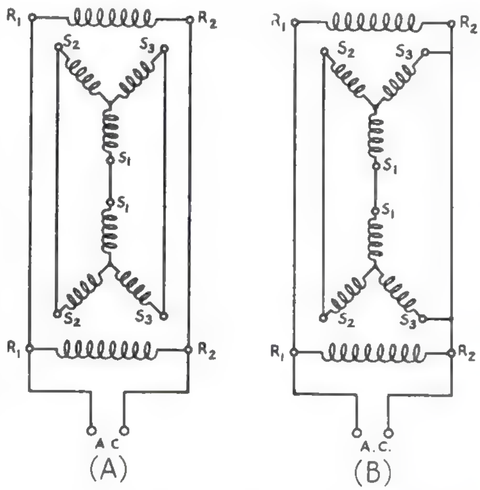


Fig. 12-42—Interconnections for synchros used for beam-direction indicators. The system at B requires only four wires instead of the usual five. With either circuit, the relative direction of rotation can be reversed by interchanging the leads to S_1 and S_2 at one of the units.

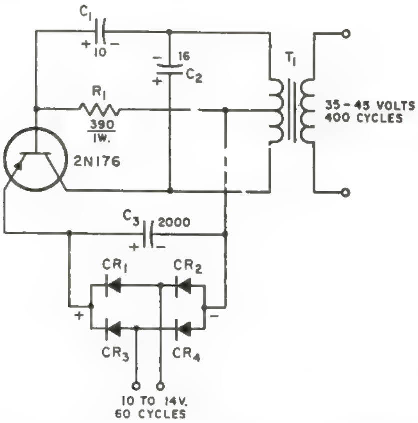


Fig. 12-43—Circuit of the 400-cycle supply (W8GZ). Capacitance is in microfarads, resistance in ohms. Capacitors are electrolytic.

C_1, C_2—50 to 150 w.v.d.c.

C_3—2000 μf. or more, 25 w.v.d.c. or more (Mallory CG452U50D1 or equivalent).

CR_1-CR_4 incl.—Bridge rectifier, 50 p.i.v., 35 volts 1.5 amps. (Mallory FW-50 bridge-rectifier unit, or four silicon diodes).

R_1—See text.

T_1—Filament transformer: 26.5 volts, c.t., 0.6 amp.; 60 cycles, secondary used as primary (Thordarson 21F27).

photographically reduced from a larger chart (some 36 inches square) centered on New York City and available for 40¢ as Chart 3042 from the Coast and Geodetic Survey. Black opaquing liquid was used to blank out the four corners, and behind this opaquing the 7-watt back-lighting bulbs are mounted. A regular shaft bearing is mounted at the center, with a counterbalanced pointer positioned over the face of the chart and the selsyn connected at the rear. The Selsyn is held in place by means of a split board—this has a cut-out hole the diameter of the body of the Selsyn, and a saw cut from one end of the board into the hole. A carriage bolt draws up on the "split" to apply pressure to the body of the Selsyn, holding it in place. The wooden frame is painted or stained to match the decor of the operating position, and is mounted on the wall over the operating position. Control cabling runs down to a suitable switch handy to the operator.

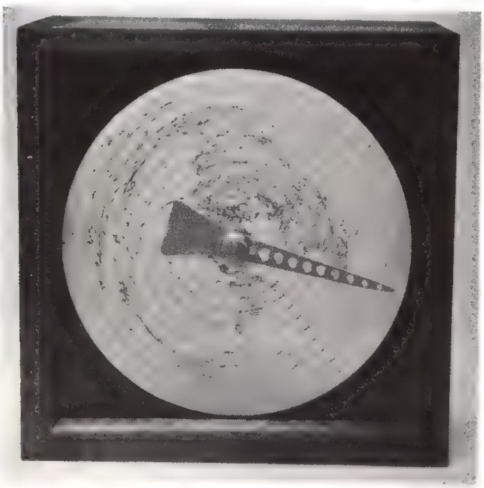


Fig. 12-44—An indirectly-lighted beam indicator using a great-circle map as the background (*W1IKE*).

The rotating-globe indicator in Fig. 12-45 uses a 7-inch globe. Locate the home town and its antipodal position, drill ⅛- or ¼-inch holes at these points, and pass a shaft through the globe, fastening the shaft to the globe with glue or suitable cement. Cut a strip of Lucite or Plexiglas to the shape shown and drill the two holes that pass the shaft. At the same time, the hairline can be scribed along the plastic, on both sides if parallax is to be avoided. The plastic is then heated until it can be bent to the shape shown in the drawing. It is then assembled with two supports on a small box that houses the Selsyn.

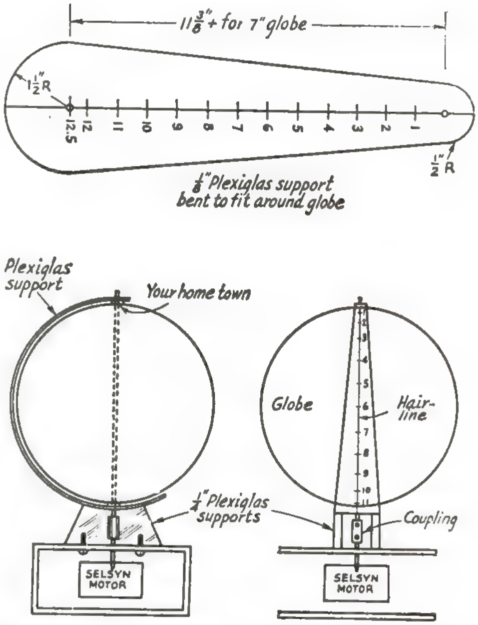


Fig. 12-45—Constructional details of the globe direction indicator. The dimensions shown are for a 7-inch globe, and should be modified for other sizes. The curved piece of Lucite is cemented to a base that is then bolted to the top of a box that houses the selsyn.

To align the indicator, point the beam due north, align the globe so that the hairline is directly over the North Pole on the globe, and tighten the coupling screws. The globe will then rotate in sync with the beam.

Finding Directions

Anyone laying out a fixed directive array does so in order to put his signal into certain parts of the world; in such cases, it is essential to be able to determine the bearings of the desired points. Too, the amateur with the rotatable directive array likes to know where to aim if he is trying to pick up certain countries. And even the amateur with the single wire is interested in the directive pattern of the lobes when the wire is operated harmonically at the higher frequencies, and often is able to vary the direction of the wire to take advantage of the lobe pattern.

FINDING DIRECTION

It is probably no news to most people nowadays that true direction from one place to another is not what it appears to be on the old Mercator school map. On such a map, if one starts "East" from central Kansas, he winds up in the neighborhood of Lisbon, Portugal. Actually, as a minute's experiment with a strip of paper on a small globe will show, a signal starting due East from Kansas never hits Europe at all but goes into the southern part of Portuguese West Africa.

If, therefore, we want to determine the direction of some distant point from our own location, the ordinary Mercator projection is utterly useless.

True bearing, however, may be found in several ways: by using a special type of world map that *does* show true direction from a specific location to other parts of the world; by working directly from a globe; or by using mathematics.

AZIMUTHAL MAPS

While the Mercator projection does not show true directions, it is possible to make up a map which will show true bearings for all parts of the world from any single point. Three such maps are reproduced in this chapter. Fig. 13-1 shows directions from Washington, D. C., Fig. 13-2 gives directions from San Francisco and Fig. 13-3 (a simplified version of the ARRL amateur radio map of the world) gives directions from the approximate center of the United States—Wichita, Kansas.

For anyone living in the immediate vicinity (within 150 miles) of any of these three reference points, the directions as taken from the maps will have a high degree of accuracy. However, one or the other of the three maps will suffice for any location in the United States for all except the most accurate work; simply pick the map whose reference point is nearest you. Greatest errors will arise when your location is to one side or the other of a line between the reference point and the destination point; if your location is near or on the resulting line, there will be little or no error.

By tracing the directional pattern of the antenna system on a sheet of tissue paper, then placing the paper over the azimuthal map with the origin of the pattern at one's location, the "coverage" of the antenna will be readily evident. This is a particularly useful stunt when a multi-lobed antenna, such as any of the long single-wire systems, is to be laid out so that the main lobes cover as many desirable directions as possible. Often a set of such patterns will be of considerable assistance in determining what length antenna to put up, as well as the direction in which it should run.

The current edition of the ARRL Amateur Radio Map of the World, entirely different in concept and design from any other radio amateurs' map, contains a wealth of information especially useful to amateurs. A special type of azimuthal projection made by Rand-McNally to ARRL specifications, it gives great circle bearings from the geographical center of the United States, as well as great circle distance measurement in miles and kilometers, within an accuracy of two per cent. The map shows principal cities of the world; local time zones; WAC divisions; more than 265 countries, indexed; and amateur prefixes throughout the world. The map is large enough to be easily readable from the operating position, 30 × 40 inches; and is printed in six colors on heavy paper. Cost is two dollars from ARRL Headquarters, 225 Main St., Newington, Conn. 06111.

The *Radio Amateur's Callbook* also includes great circle maps and tables, and another *Callbook* publication, *The Radio Amateur's World Atlas* (price, $1.50), features a polar-projection world map, maps of the continents, and world amateur prefixes. The maps are in color.

WORKING FROM A GLOBE

Entirely satisfactory bearings for beam purposes can be taken from an ordinary globe with nothing more complicated than a small school protractor of the type available in any school-

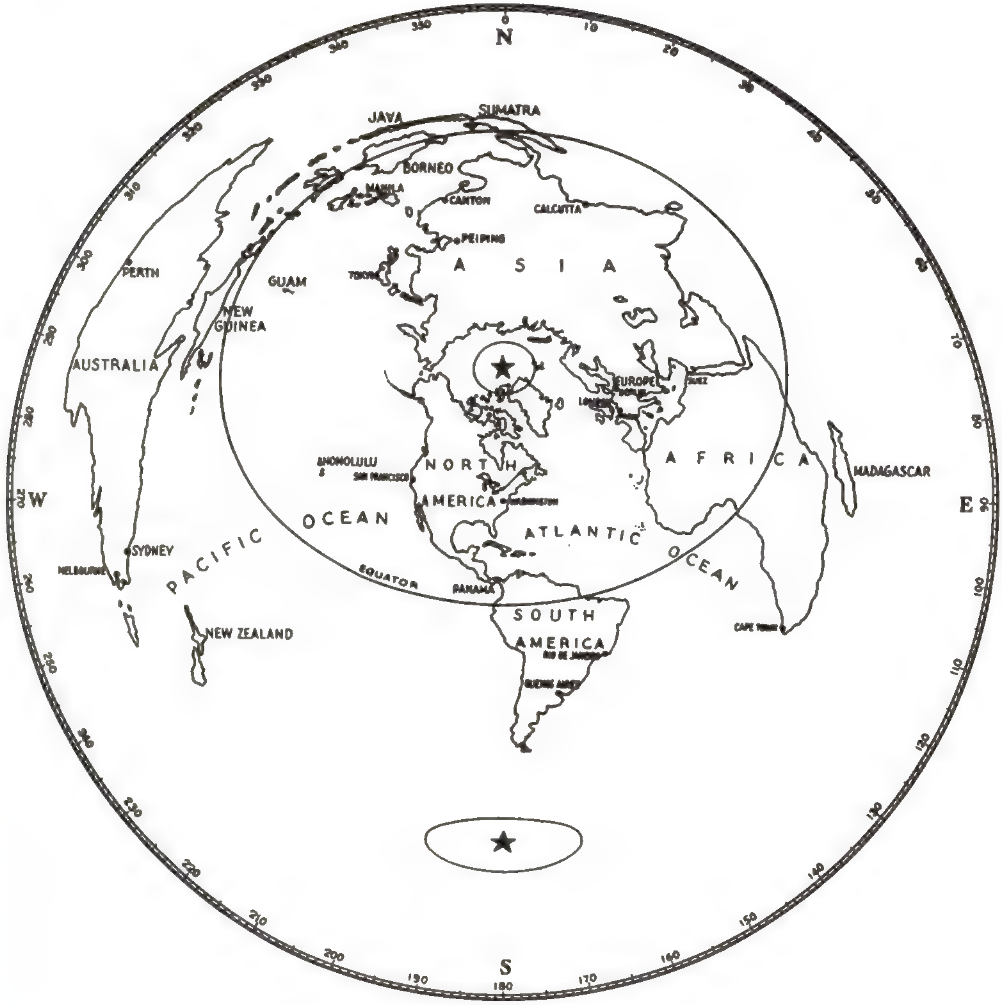


Fig. 13-1—Azimuthal map centered on Washington, D. C.

supply or stationery store, as illustrated in Fig. 13-4. For best results, however, the globe should be at least eight inches in diameter.

From a piece of thin paper, cut out a small circle—something like a three-inch circle for use with an eight-inch globe. Put a pin through the center and draw a straight line from the center to any point on the circumference. Now, put the paper circle on the globe, sticking the center pin into your location. Using the edge of a sheet of plain paper as a straightedge, line up the straight line on your paper circle so that it points North; this is done by laying the straightedge against the center pin and running it up to the North Pole at the top of the globe, then turning the paper circle until the straight line on it coincides with the straightedge. When you have done this, stick another pin through the paper circle into the globe to hold it in position with this line pointing North.

Now all you have to do is to use your paper straightedge from the center pin to such points as you wish, drawing short lines on your paper circle and labeling them as required. These lines may be extended later to the periphery of the circle.

With your protractor it is now a simple matter to determine the bearing, in degrees from North of any of the points.

If your problem is to lay out a long wire to best advantage, make a diagram from the data in Chapter Two, showing the angular direction of the lobes, and superimpose this on your direction chart, adjusting it until the theoretical power lobes seem to take in the points in which you are interested. The direction of the wire can then be determined with the protractor.

DIRECTION AND DISTANCE BY TRIGONOMETRY

The method to be described will give the bearing (and distance) as accurately as one

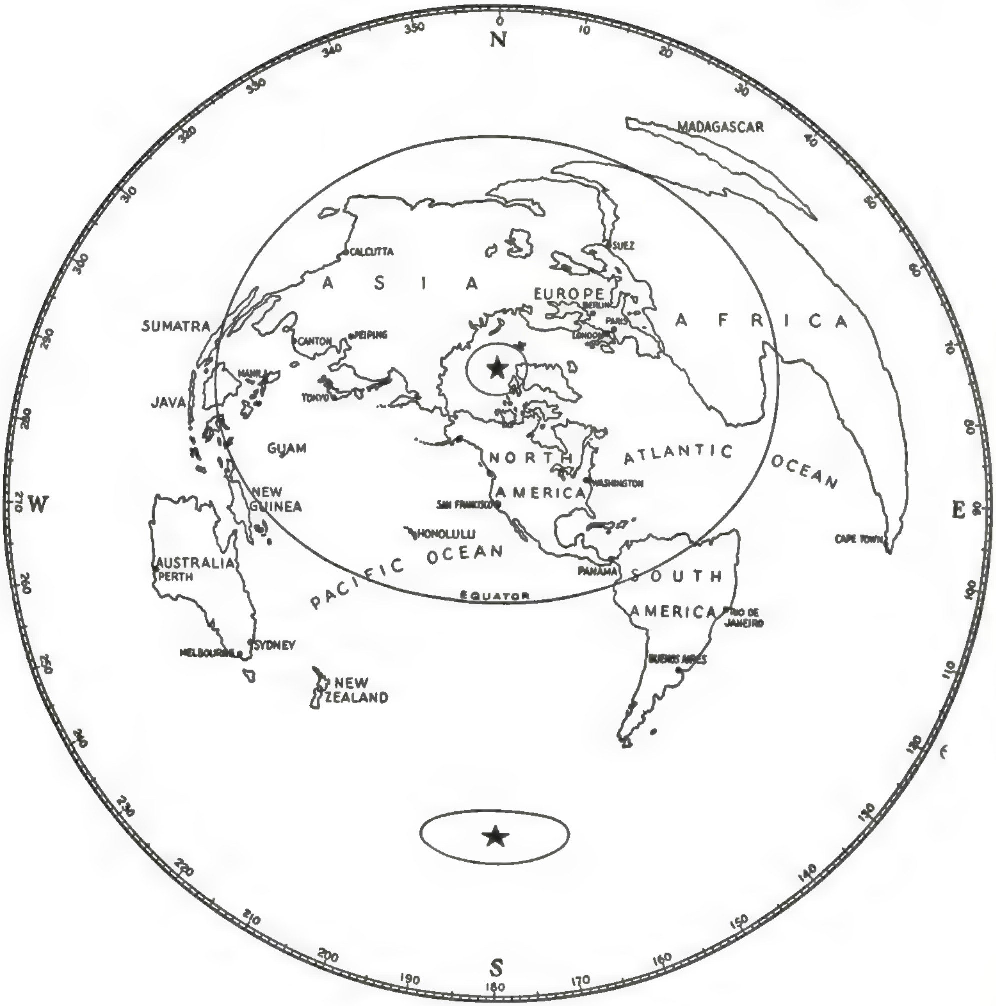


Fig. 13-2—Azimuthal map centered on San Francisco, Calif.

cares to compute them. All that is required is a table of latitude and longitude, such as may be found in the *World Almanac*, and a set of trigonometry tables. For most purposes, the latitude and longitude can be taken from maps of the areas in question. Two formulas are used:

$\cos D = \sin A \sin B + \cos A \cos B \cos L \ldots (1)$

and $\sin C = \cos B \csc D \sin L \ldots\ldots\ldots (2)$

where A = *your* latitude in degrees

B = the other location in degrees (positive for N latitude, negative for S latitude)

L = longitude difference between you and the other location

C = the direction of the other location from yours, in degrees East or West from North or South

D = distance along path, in nautical miles or minutes of arc (1 min. = 1 nautical mile = 1.15078 statute miles)

The following example will show how the formulae are used. To find the bearing and distance of Cairo from Chicago:

From the *Almanac* or map:

Chicago = 41° 52′ N, 87° 38′ W

Cairo = 30° 00′ N, 31° 14′ E

$A = 41° 52'$

$B = 30°$

$L = 87° 38' + 31° 14' = 118° 52'$

Substituting in (1):

$$\cos D = \sin(41°\ 52') \times \sin(30°) + \cos(41°\ 52') \times \cos(30°) \times \cos(118°\ 52')$$
$$= .6774 \times .5 + .7447 \times .8660 \times (-.4828)$$
$$= .3387 - .3114$$
$$= .0273$$
$$D = 88°\ 26' = (88 \times 60 + 26 = 5280 + 26 \text{ min.}$$
$$= 5306 \text{ nautical miles}$$
$$= 5306 \times 1.15078 = 6106 \text{ statute miles}$$

Substituting in (2):

$$\sin C = \cos(30°) \times \csc(88°\ 26') \times \sin(118°\ 52')$$

Fig. 13-3—Azimuthal map centered on Wichita, Kansas.

$$= .8660 \times \frac{1}{\sin (88^\circ\ 26')} \times .8758 = \frac{.7584}{.9996}$$
$$= .7587$$
$$C = 49^\circ\ 21'$$

Thus the true bearing of Cairo is 49° 21′ from Chicago and, by inspection of a globe, this direction is seen to be *East* of North. With the antenna in Chicago pointed 49° 21′ East of North, a maximum signal should be pumped into Cairo, 6106 miles away. Similarly, the direction and range between any two points on the earth can be computed.

DETERMINING TRUE NORTH

Determining the direction of distant points is of little use to the amateur erecting a directive array unless he can put up the array itself in the desired direction. This, in turn, demands a knowledge of the direction of *true* North (as against magnetic North), since all our directions from globe or map are worked in terms of true North.

A number of ways may be available to the amateur for determining true North from his location. Frequently, the streets of a city or town are laid out, quite accurately, in North-South and East-West directions. A visit to the office of your city engineer will enable you to determine whether or not this is the case for the street in front of or parallelling your own lot. Or from such a visit it is often possible to locate some landmark, such as factory chimney or church spire, which lies true North with respect to your house.

If you cannot get true North by such means, three other methods are available: compass, Pole Star and Sun.

By Compass: Get as large a compass as you can; it is difficult, though not impossible, to get

Fig. 13-4—A direction indicator made from a semicircle of thin metal can be fitted easily to a small globe. Pins at the ends permit fastening one end to the home location, the other to the antipodes. The paper scale is marked in miles to show approximate distances (12,500 miles to the semicircle).

satisfactory results with the "pocket" type. In any event, the compass *must* have not more than two degrees per division.

It must be remembered that the compass points to *magnetic* North, not true North. The amount by which magnetic North differs from true North in a particular location is known as *variation.* Your local weather bureau or city engineer's office can tell you the magnetic variation for your locality. When correcting your "compass North," do so *opposite* to the direction of the variation. For instance, if the variation for your locality is 12 degrees West (meaning that the compass points 12 degrees West of North) then true North is found by counting 12 degrees *East* of North as shown on the compass.

When taking the bearing, make sure that the compass is located well away from ironwork, fencing, pipes, etc. Place the instrument on a wooden tripod or support of some sort, at a convenient height as near eye level as possible. Make yourself a sighting stick from a flat stick about two feet long with a nail driven upright in each end (for use as "sights") and then, after the needle of the compass has settled down, carefully lay this stick across the face of the compass—with the necessary allowance for variation—to line it up on true North. *Be sure you apply the variation correctly.*

This same sighting-stick-and-compass rig can also be used in laying out directions for supporting poles for antennas in other directions—provided, of course, that the compass dial is graduated in degrees.

By the Pole Star: Many amateurs use the Pole Star in determining the direction of true North. An advantage is that the Pole Star bears true North, so that no corrections are necessary. Disadvantages are that some people have difficulty identifying the Pole Star, which is none too bright at best, and that because of its comparatively high angle above the horizon it is not always easy to "sight" on it accurately.

In any event, it is a handy check on the direction secured by other means.

By the Sun: With some slight preparation, the sun can easily be used for determination of true North. One of the most satisfactory methods is described below. The method is based on the fact that exactly at noon, local time, the Sun bears due South, so that at that time the shadow of a vertical stick or rod will bear North. The resulting shadow direction, incidentally, is *true* North.

Two corrections to your standard time must be made to determine the exact moment of local true noon.

The first is a longitude correction. Standard Time is time at some particular meridian of longitude: EST is based on the 75th meridian; CST on the 90th meridian; MST on the 105th meridian; and PST on the 120th meridian. From an atlas, determine the difference between your own longitude and the longitude of your time meridian. Getting this to the nearest 15 minutes of longitude is close enough. Example: West Hartford, Conn., which runs on 75th meridian time (EST) is at 72° 45′ longitude, or a difference of 2° 15′. Now, for each 15′ of longitude, figure 1 minute of time; thus 2° 15′ is equivalent to 9 miuutes of time (there are 60 "angle" minutes to a degree, so that each degree of longitude equals 4 minutes of time). *Subtract* this correction from noon if you are

TABLE 13-1

Apply to Clock Time as Indicated by the Sign, To Get Time of True Noon

Jan. 1	+ 4 min.	July 10	+ 5 min.
10	+ 8 "	20	+ 6 "
20	+ 11 "	30	+ 6 "
30	+ 13 "		
Feb. 10	+ 14 "	Aug. 10	+ 5 "
20	+ 14 "	20	+ 3 "
28	+ 13 "	30	+ 1 "
Mar. 10	+ 10 "	Sept. 10	− 3 "
20	+ 8 "	20	− 7 "
30	+ 4 "	30	− 10 "
Apr. 10	+ 1 "	Oct. 10	− 13 "
20	− 1 "	20	− 15 "
30	− 3 "	30	− 16 "
May 10	− 4 "	Nov. 10	− 16 "
20	− 4 "	20	− 14 "
30	− 3 "	30	− 11 "
June 10	− 1 "	Dec. 10	− 7 "
20	+ 1 "	20	− 2 "
30	+ 4 "	30	+ 3 "

East of your time meridian; *add* it if you are *West*.

To the resulting time, apply a further correction for the date from Table 13-I. The resulting time is the time, by Standard Time, when it will be true noon at your location. Put up your vertical stick (use a plumb bob to make sure it is actually vertical), check your watch with Standard Time and, at the time indicated from your calculations, mark the position of the shadow. That is true North.

(In the case of West Hartford, if we wanted correct time for true noon on October 20: First, subtracting the longitude correction—because we are East of the time meridian—we get 11:51 A.M.; then, applying the further correction of —15 minutes, we get 11:36 A.M. as the time of true noon at West Hartford on October 20.)

Receiving Antennas

For the amateur engaged in traffic work or general rag-chewing, where operating convenience overshadows any other factor, a wire antenna of any convenient length is normally quite satisfactory for reception. Such a separate receiving antenna should preferably be as far removed from the transmitting antenna as possible, so that a minimum of energy from the transmitter will be picked up.

If a short wire picks up too much "man-made noise" in the form of hash and interference from motors and other electrical appliances, it is advisable to use a half-wave antenna cut for the frequency band most often used. It should be used with a low-impedance line connected to its center, just as in the transmitting case, for maximum signal transfer to the receiver. The folded dipole with 300-ohm feeders is a very good receiving antenna. By placing the antenna at a point removed from the source of noise, the noise will be reduced and, since the line is balanced, the low-impedance line will run back through the noisy area but will not pick up any noise. This is shown in Fig. 14-1. Low-impedance line will not in itself reduce noise, but it allows the antenna to be placed at a distance and still feed its signal back to the receiver. If a half-wave antenna is used, it should be placed at right angles to the transmitting antenna and as far away as possible.

RECEIVING WITH DIRECTIVE ANTENNAS

The amateur fortunate enough to own a directive antenna should always use it for receiving as well as for transmitting, instead of using a separate receiving antenna. When separate antennas are used, a signal might be heard from a direction in which the transmitting antenna has a null, and all of the calling one can do won't result in a QSO if there is no signal going out in the right direction. With rotary antennas, one can rotate the beam until the received signal peaks up, and the operator can then feel assured that his antenna is aimed at the distant station.

Using the same antenna for receiving and transmitting requires switching of some kind. This can be as simple (and inconvenient) as a hand-operated switch. A better device is an electrically-actuated antenna changeover relay. If break-in operation is desired, an electronic "t.r." (transmit-receive) switch can be used. The construction of t.r. switches has been described in *QST* and *The Radio Amateur's Handbook*, and there are commercial models available.

If two or more directive systems are available, some provision should be made for quick switching from one to the other or others, so that when listening one can identify the direction of the signal even before the station signs. Even with antennas with no marked directional characteristics, switching will show which one yields the loudest signal and hence should be used for transmitting as well. The coupling to the receiver should duplicate the transmitter coupling, however, to insure that the antenna system works the same for transmitting and receiving.

Although some stations use a hand-operated switch to change the antenna from the transmitter to the receiver and back again, it is far more satisfactory to use antenna change-over relay. These are made with good insulation and work from the 115-volt a.c. line, so that it is only necessary to connect the relay coil across the primary of a plate transformer so that every time the transmitter is switched on the antenna relay operates and switches the antenna from the receiver to the transmitter.

The change-over relay (or switch) does not

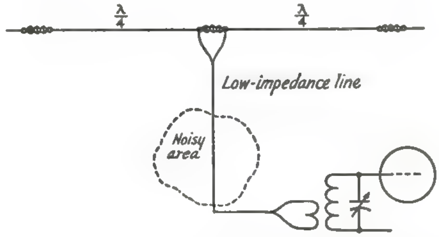


Fig. 14-1—The doublet antenna with low-impedance line is used for noise reduction by placing the antenna outside or away from the noisy area and running a low-impedance line to it. The line itself doesn't reduce the noise but, because it has little pick-up itself, it enables the signal from the antenna to be fed to the receiver through an interference-generating area.

necessarily have to be connected in the antenna feeder. If an antenna tuning unit is used, coupled to the transmitter through a length of coax line, the relay can be placed in the coax and the receiver will then be able to take advantage of the matching afforded by the tuning unit. The various ways in which the change-over relay (or switch) can be used are shown in Fig. 14-2.

Using the same antenna tuning unit for both transmitting and receiving is advantageous in that it reduces the amount of equipment needed; furthermore, the greatest response to incoming signals is automatically secured at and near the frequency to which the transmitter is tuned. However, the separate tuner is more convenient in that it is possible to listen on other frequencies, and on other bands, without loss of signal strength and without disturbing the tuning for transmission.

The circuits of Figs. 14-2E and 14-2F show no coupling device to the receiver. In many cases this is perfectly satisfactory, particularly at frequencies below 10 Mc., but additional signal strength can often be obtained by matching the receiver to the line. Methods for doing this are shown in Fig. 14-3. The networks should be located fairly close to the receiver, in a position convenient for occasional adjustment by the operator.

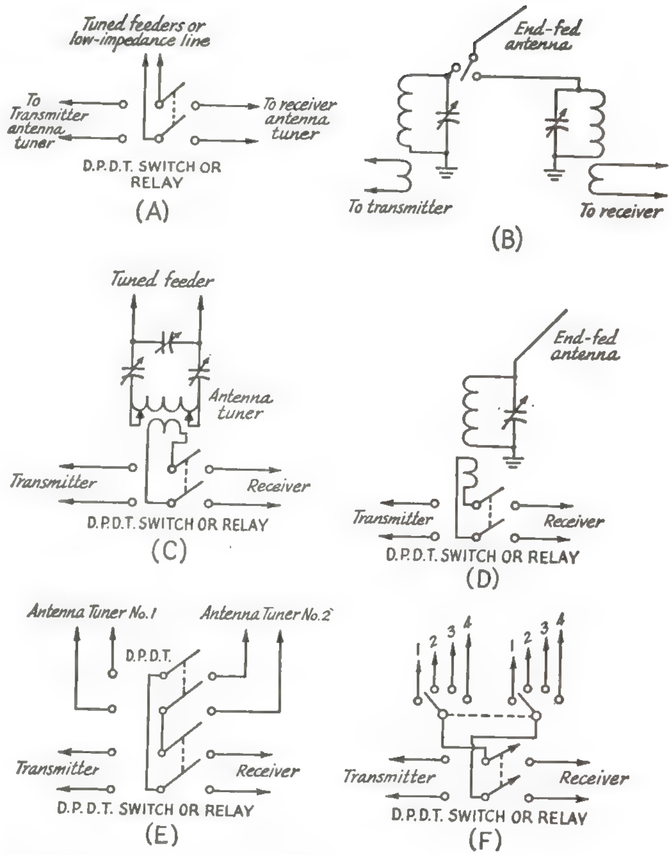


Fig. 14-2—Antenna switching systems. A—For tuned or untuned lines with separate antenna tuners. B—For voltage-fed antenna. A series-tuned circuit will be required with some antenna lengths. C—For tuned line wtih single tuner. D—For voltage-fed antenna with single tuner. See B above. E—For two tuned-line antennas with tuner for each antenna or for low-impedance lines. F—For several two-wire lines.

In some instances it will be found that a rotary beam used for receiving does not show the rejection to unwanted signals off the side that might be expected from such an antenna. In most of these cases, it will be found that the receiver is responding to parallel components on the transmission line (see Chapter Four). To make sure that the receiver is not at fault, connect several feet of the same type of line to the receiver, to see if signals can be picked up on the line alone. If a balanced line such as 300-ohm Twin-Lead is used, connect it to the two antenna terminals of the receiver, leaving the ground post open. If a shielded two-wire line, or two pieces of coaxial line, is used, connect the inner conductors to the antenna terminals and the outer conductor(s) to the ground post. With single coaxial line, connect the inner conductor to one antenna terminal and the outer conductor to ground *and* the other antenna terminal. No signals should be heard, even with the receiver running at maximum gain. When a two-wire (balanced) line is used not all receivers will pass this test successfully, because of capacitive coupling between the antenna coil in the receiver's first tuned circuit and the tuned-circuit coil itself. A receiver having a coaxial antenna fitting usually will do much better, because the parallel signal component (which is on the *outside* of the line) will be shunted to ground, with proper receiver construction.

If signals are heard, an antenna coupler as shown in Fig. 14-3C will help to eliminate them if a single length of coaxial line is run from L_5 to the receiver and connected as described above for the test with single coaxial line. However, the receiver must be capable of passing the test for single coaxial line.

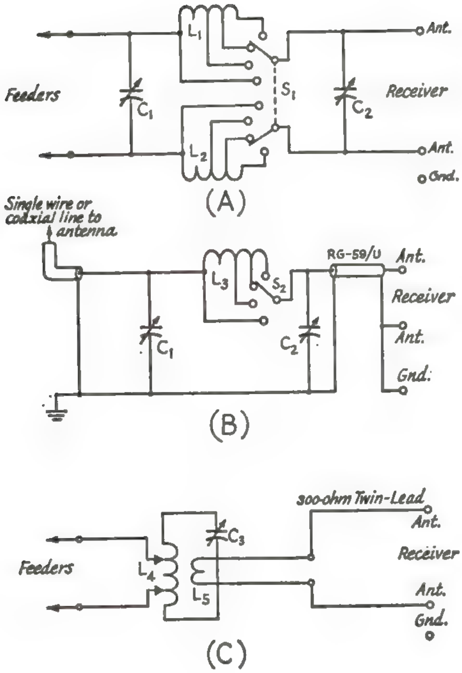


Fig. 14-3—Three types of circuits for coupling the antenna to the receiver. A—Balanced pi-section network for use with balanced lines of any kind. B—Single-ended pi-section network for use with single wires or coaxial line. C—Tuned circuit for use with any balanced line.

C_1, C_2, C_3—100-pf. midget variable.

L_1, L_2, L_3—30 turns No. 28 d.c.c. close-wound on ½-inch diameter form, tapped at 2½, 6½ and 14½ turns.

L_4—Proportioned to resonate with C_3 in the desired band.

L_5—2- or 3-turn swinging link at center of L_4.

S_1—2-circuit 5-position single-section ceramic water switch.

S_2—1-circuit 5-position single-section ceramic wafer switch.

GROUNDS

Most modern receivers do not require an external ground, since they are grounded through the power-supply line by the capacitance of the transformer windings. However, in some instances a direct ground to the receiver will boost the signal pickup. The direct ground can be made by running a wire to a radiator or water pipe or directly to a pipe in the ground.

LOOP ANTENNAS

Loop antennas are sometimes useful for regular reception on the 7- and 3.5-Mc. bands. In the usual location, the pick-up inside a house at the high frequencies is not too good, and metallic objects in the vicinity, such as house wiring and transmitter control wires, tend to confuse the directional characteristics of the loop. In addition, most man-made electrical noise is carried by power wiring and is most bothersome in the immediate vicinity of such wiring, so that a loop indoors near house wiring may pick up more noise than an outside antenna, although sometimes such noise can be "nulled" out if its direction is well defined. In general, the directional properties of even a good loop are not too useful on signals that have been propagated through the ionosphere because of the random polarization and wave angle of such signals.

However, a loop can be helpful at times in rejecting unwanted signals, and two types are shown in Fig. 14-4. The loop at A is a little less elaborate than that of B and will not give quite as good results, but it is still useful in rejecting signals.

In either case, the loop should be made from not more than about 20 feet of insulated wire, formed in a square of from 12 to 20 inches on a side or a circle of from 12 to 20 inches in diameter. It can be mounted on the face of a sheet of plywood, and it should be hinged in both the vertical and horizontal planes, so that it can be pointed in any direction. The tuning capacitor

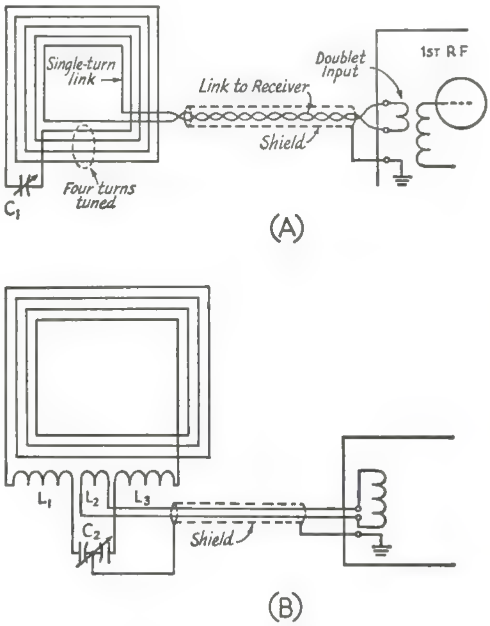


Fig. 14-4—Two types of loop antennas for 3.5-Mc. reception. The loops are made of not more than about 20 feet of No. 18 stranded wire formed in a circle or square of 3, 4 or 5 turns.

C_1—150-pf. variable.

C_2—100 pf. each section.

L_1, L_3—17 turns No. 20, wound on 1½-inch diameter form and spaced to occupy ¾-inch winding length.

L_2—4-turn link winding, 1½-inch diameter, movable between L_1 and L_3.

is adjusted for maximum signal and, in the case of the loop of Fig. 14-4B, the coupling between L_2 and L_1L_3 should also be adjusted for best results.

Although the signal strength with a loop antenna is smaller than from a regular antenna, the sharp null of a properly-constructed loop may make reception possible through noise or interference which could not be carried on with a conventional antenna.

Mobile Antennas

The Whip Antenna

EQUIVALENT CIRCUIT

For mobile operation in the range between 1.8 and 30 Mc., the vertical whip antenna is the most-frequently used. The length is usually limited to a dimension that will resonate as a quarter-wave antenna in the 10-meter band, since longer whips present mechanical difficulties. This length is approximately 8 feet. The car body acts principally as a ground plane at the higher frequencies, and as a capacitance to ground at the lower frequencies.

With the whip length adjusted to resonance in the 10-meter band, the impedance at the feed point, *X*, Fig. 15-1, will appear as a pure resistance at the resonant frequency. This resistance will be composed almost entirely of radiation resistance, and the efficiency will be high. However, at frequencies lower than the resonant frequency, the antenna will show an increasingly higher capacitive reactance and a decreasingly lower radiation resistance.

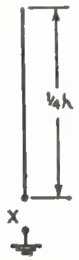


Fig. 15-1—The quarter-wave whip at resonance will show a pure resistance at the feed point, X.

The equivalent circuit is shown in Fig. 15-2. For the average 8-ft. whip, the reactance of C_A may range from about 150 ohms at 21 Mc. to as high as 8000 ohms at 1.8 Mc., while the radiation resistance R_R varies from about 15 ohms at 21 Mc. to as low as 0.1 ohm at 1.8 Mc.

For an antenna less than 0.1 wavelength long, the approximate radiation resistance may be determined from the following:

$$R_R = 273\,(lf)^2 \times 10^{-8},$$

where l is the length of the whip in inches, and f is the frequency in megacycles.

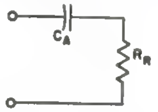


Fig. 15-2—At frequencies below the resonant frequency, the whip antenna will show capacitive reactance as well as resistance. R_R is the radiation resistance, and C_A represents the capacitive reactance.

Since the resistance is low, considerable current must flow in the circuit if any appreciable power is to be dissipated as radiation in the resistance R_R. Yet it is apparent that little current can be made to flow in the circuit so long as the comparatively high series reactance remains.

Eliminating Reactance

The capacitive reactance can be canceled out by connecting an equivalent inductive reactance, L_L, in series, as shown in Fig. 15-3, thus tuning the system to resonance.

Fig. 15-3—The capacitive reactance at frequencies lower than the resonant frequency of the whip can be canceled out by adding an equivalent inductive reactance in the form of a loading coil in series with the antenna.

Unfortunately, all coils have resistance, and this will be added in series, as indicated at R_C in Fig. 15-4. While a large coil may radiate some energy, thus adding to the radiation resistance, the latter will usually be small compared to the loss resistance introduced. However, adding the coil makes it possible to feed power to the circuit.

Ground Loss

Another element in the circuit dissipating power is the ground-loss resistance. Funda-

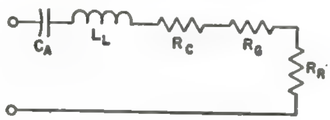


Fig. 15-4—Equivalent circuit of a loaded whip antenna. C_A represents the capacitive reactance of the antenna, L_L an equivalent inductive reactance. R_G the ground-loss resistance, and R_R the radiation resistance.

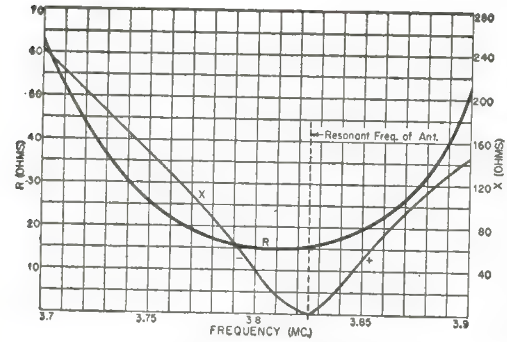


Fig. 15-5—Typical reactance and total-resistance curves of an unloaded 8-ft. whip over the 75-meter band, measured through a half-wave line of RG-8/U.

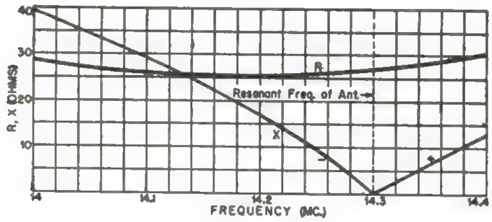


Fig. 15-6—Typical reactance and total-resistance curves of an unloaded 8-ft. whip over the 20-meter band, measured through a half-wave line of RG-8/U.

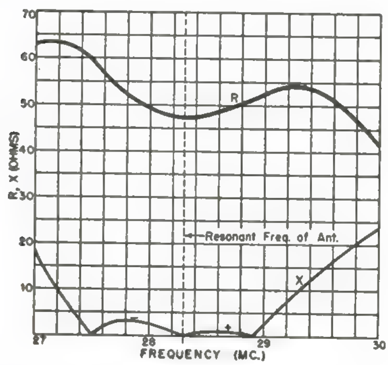


Fig. 15-7—Typical reactance and total-resistance curves of an 8-ft. whip over the 10-meter band, measured through an RG-8/U line three wavelengths long. The measured reactance was positive in the region from 27 to 27.5 Mc. and negative between 28.9 and 30 Mc., because of effects introduced by the measuring line. The actual antenna reactance would not exhibit such a change in polarity.

mentally, this is related to the nature of the soil in the area under the antenna. Little information is available on the average values of resistance to be expected in practice, but some measurements have shown that it may amount to as much as 10 or 12 ohms at 4 Mc. At the lower frequencies, it may constitute the major circuit resistance.

Fig. 15-4 shows the circuit including all of the elements mentioned above. Assuming C_A lossless and the loss resistance of the coil to be represented by R_C, it is seen that the power output of the transmitter is divided among three resistances—R_C, R_G, and R_R. Only the power dissipated in R_R is radiated. The power developed in R_C and R_G is dissipated in heat. Therefore, it is important that the latter two resistances be minimized.

Figs. 15-5, 15-6, and 15-7 show typical curve for an 8-ft. whip over the 3.8-, 14- and 28-Mc bands. Curves for 7 and 21 Mc. will lie intermediate to those shown. Resistances are total, including coil- and ground-loss resistances.

MINIMIZING LOSSES

There is little that can be done about the nature of the soil. However, poor electrical contact between large surfaces of the car body, and especially between the point where the feed line is grounded and the rest of the body, can add materially to the ground-loss resistance. For example, the feed line, which should be grounded as close to the base of the antenna as possible, may be connected to the bumper, while the bumper may have poor contact with the rest of the body because of rust or paint.

Loading Coils

The capacitance of a vertical antenna shorter than a quarter wavelength is given by

$$C_A = \frac{17L}{\left[\left(\log_e \frac{24L}{D}\right) - 1\right]\left[1 - \left(\frac{fL}{246}\right)^2\right]}$$

where

C_A = capacitance of antenna in pf.
L = antenna height in feet
D = diameter of radiator in inches
f = operating frequency in Mc.

$$\log_e \frac{24L}{D} = 2.3 \log_{10} \frac{24L}{D}$$

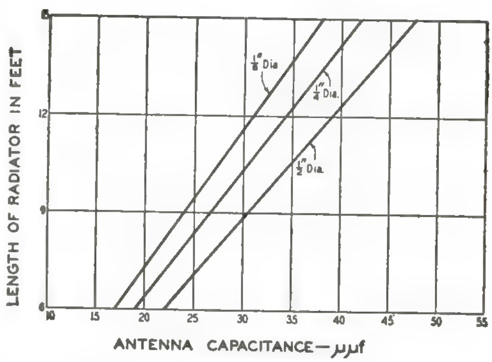


Fig. 15-8—Graph showing the approximate capacitance of short vertical antennas for various diameters and lengths. These values should be approximately halved for a center-loaded antenna.

Table 15-I
Approximate Values for 8-ft. Mobile Whip

f_{kc}	Loading $L_{\mu h}$	R_C (Q50) Ohms	R_C (Q300) Ohms	R_R Ohms	Feed R* Ohms	Matching $L_{\mu h}$.**
Base Loading						
1800	345	77	13	0.1	23	3
3800	77	37	6.1	0.35	16	1.2
7200	20	18	3	1.35	15	0.6
14,200	4.5	7.7	1.3	5.7	12	0.28
21,250	1.25	3.4	0.5	14.8	16	0.28
29,000	....	....	....	....	36	0.23
Center Loading						
1800	700	158	23	0.2	34	3.7
3800	150	72	12	0.8	22	1.4
7200	40	36	6	3	19	0.7
14,200	8.6	15	2.5	11	19	0.35
21,250	2.5	6.6	1.1	27	20	0.29

R_C = Loading-coil resistance; R_R = Radiation resistance.
* Assuming loading coil Q = 300, and including estimated ground-loss resistance.
** For matching given feed resistance to 52 ohms.
Suggested coil dimensions for the required loading inductances are shown in Table 15-II.

The graph of Fig. 15-8 shows the approximate capacitance of whip antennas of various average diameters and lengths. For 1.8, 4 and 7 Mc., the loading-coil inductance required (when the loading coil is at the base) will be approximately the inductance required to resonate in the desired band with the whip capacitance taken from the graph. For 14 and 21 Mc., this rough calculation will give more than the required inductance, but it will serve as a starting point for the final experimental adjustment that must always be made.

Table 15-I shows the approximate loading-coil inductance for the various amateur bands. Also shown in the table are approximate values of radiation resistance to be expected with an 8-ft. whip, and the resistances of loading coils—one group having a Q of 50, the other a Q of 300. A comparison of radiation and coil resistances will show the importance of reducing the coil resistance to a minimum, especially on the three lower-frequency bands.

To minimize loading-coil loss, the coil should have a high ratio of reactance to resistance, i.e., high Q. A 4-Mc. loading coil wound with small wire on a small-diameter solid form of poor quality, and enclosed in a metal protector, may have a Q as low as 50, with a resistance of 50 ohms or more. High-Q coils require a large conductor, "air-wound" construction, turns spaced, the best insulating material available, a diameter not less than half the length of the coil (not always mechanically feasible), and a minimum of metal in the field. Such a coil for 4 Mc. may show a Q of 300 or more, with a resistance of 12 ohms or less. This reduction in loading-coil resistance may be equivalent to increasing the transmitter power by 3 times or more. Most low-loss transmitter plug-in coils of 100-watt size or larger, commercially produced, show a Q of this order. Where large inductance values are required, lengths of low-loss space-wound coils are available (B & W or Airdux). If a weather shield is to be used, it should be of plastic, not metal.

Several manufacturers produce a series of loading coils for whip antennas. They come complete with suitable fittings for mounting standard types of antennas.

Center Loading

The radiation resistance of the whip antenna can be approximately doubled, at the lower frequencies, by placing the loading coil at the center of the whip, rather than at the base, as shown in Fig. 15-9. (The optimum position varies with ground resistance. The center is optimum for average ground resistance.) However, because the capacitance of only that portion of the whip above the coil appears across the coil, the inductance of the loading coil must be approxi-

Table 15-II
Suggested Loading-Coil Dimensions

Req'd $L_{\mu h}$.	Turns	Wire Size	Diam. In.	Length In.	Form or B & W Type
700	190	22	3	10	Polystyrene
345	135	18	3	10	Polystyrene
150	100	16	2½	10	Polystyrene
77	75	14	2½	10	Polystyrene
77	29	12	5	4¼	160T
40	28	16	2½	2	80B less 7 t.
40	34	12	2½	4¼	80T
20	17	16	2½	1¼	80B less 18 t.
20	22	12	2½	2¾	80T less 12 t.
8.6	16	14	2	2	40B less 4 t.
8.6	15	12	2½	3	40T less 5 t.
4.5	10	14	2	1¼	40B less 10 t.
4.5	12	12	2½	4	40T
2.5	8	12	2	2	15B
2.5	8	6	2⅜	4½	15T
1.25	6	12	1¾	2	10B
1.25	6	6	2⅜	4½	10T

mately doubled over the value required at the base to tune the system to resonance. For a coil of the same Q, the coil resistance will also be doubled. But, even if this is the case, center

Fig. 15-9—Placing the loading coil at the center of the whip antenna, instead of at the base, increases the radiation resistance, although a larger coil must be used.

loading represents a gain in antenna efficiency, especially at the lower frequencies. This is because the ground-loss resistance remains the same, and the increased radiation resistance becomes a larger portion of the total circuit resistance, even though the coil resistance is also increased. However, as turns are added to a loading coil (other factors being equal) the inductance (and therefore the reactance) increases at a faster rate than the resistance, and the larger coil will usually have the higher Q.

Fig. 15-10—The top-loaded 4-Mc. antenna used by W6SCX. The loading coil may be a space-wound transmitting coil mounted on a heavy insulating rod into which the two antenna sections are threaded.

Top Loading Capacitance

Since the coil resistance varies with the inductance of the loading coil, the coil resistance can be reduced by reducing the number of turns. This can be done, while still maintaining resonance, by adding capacitance to the portion of the antenna above the coil. This capacitance can be provided by attaching a capacitive surface as high up on the antenna as is mechanically feasible. Capacitive "hats," as they are usually called, may consists of a light-weight metal ball, cylinder, disk or wheel structure. The latter type is shown in Fig. 15-10. Fig. 2-84 (Chapter Two) shows the approximate added capacitance to be expected from top-loading devices of various forms and dimensions. This top-loading capacitance should be added to the capacitance of the portion of the whip above the loading coil (from Fig. 15-8) in determining the approximate inductance of the loading coil.

When center loading is used, the amount of capacitance to be added to permit the use of the same loading inductance as is required for base loading is not as great. Center loading should be seriously considered, since the total gain to be made by moving the coil to the center of the antenna may be quite marked.

Antenna Mounting

The whip will radiate best when its base is mounted as high as possible on the car, and as far away as possible from vertical surfaces of the car. Cowl and rear-deck mountings are preferable to bumper-level mounting.

FEEDING THE WHIP ANTENNA

Shunt Feed

It is usually found most convenient to feed the whip antenna with coax line. Unless very low-Q (high-resistance) loading coils are used, the feedpoint impedance will always be lower than 52 ohms, the impedance of the commonly-used coax line, RG-8/U or RG-58/U. Since the length of the transmission line will seldom exceed 10 or 12 feet, the losses involved will be negligible, even with a fairly-high standing-wave ratio. However, unless a line of this length is made reasonably flat, difficulty may be encountered in obtaining sufficient coupling to load the transmitter output stage, especially if the tank circuit is one having untuned-link output.

One method of obtaining a match is shown in Fig. 15-11. A small inductance, L_M, is inserted at the base of the antenna, the loading-coil inductance being reduced correspondingly to maintain resonance. The line is then tapped on the matching coil at a point where the desired loading is obtained. The approximate inductive reactance that must be added in series with the

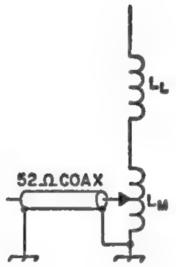


Fig. 15-11—A method of matching the loaded whip to 52-ohm coax cable. L_L is the loading coil and L_M the matching coil.

base of the antenna and ground (across the coax line) is given by:

$$X_{L_M} = \frac{Z_L R_A}{\sqrt{R_A (Z_L - R_A)}}$$

where Z_L is the characteristic impedance of the coax line, and R_A is the total resistance of the antenna circuit, including coil and ground-loss resistances.

Table 15-I shows the approximate inductance to be used between the line tap and ground for the various amateur bands. It is advisable to make the experimental matching-coil inductance somewhat larger than the value shown in the table, to make provision for varying the inductance for the best match.

For operation in the bands from 28 to 1.8 Mc., the whip should first be resonated at 28 Mc. with the matching coil inserted, but the line disconnected, using a grid-dip oscillator coupled to the matching coil. Then the line should be attached, and the tap adjusted to make the antenna system "look like" 52 ohms. This is best done with an s.w.r. bridge although, as stated earlier, the loss in a short length of coax will be inconsequential even if the s.w.r. is fairly high. If no bridge is available it will suffice to find a tap position that will permit the transmitter's final stage to be loaded to normal input; in the case of link-coupled circuits, select the tap position that results in the least detuning of the final tank capacitor as compared with its setting without load. If it is found, when using the s.w.r. bridge, that the s.w.r. cannot be brought down to 1 to 1, even though there appears to be sufficient range of tap adjustment, the antenna should be readjusted for resonance, since a mistuned antenna cannot be matched accurately.

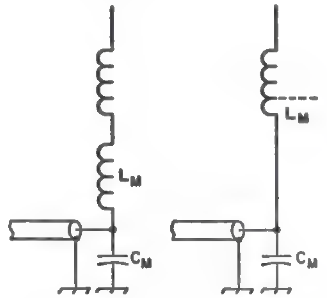


Fig. 15-13—A whip antenna may also be matched to coax line by means of an *L* network. The inductive reactance of the *L* network can be combined in the loading coil, as indicated at the right.

The same procedure should be followed for each of the other bands, first resonating by adjusting the loading coil with the g.d.o. coupled to the matching coil.

After the position of the matching tap has been found, the size of the matching coil can be reduced to just that portion between the tap and ground, if desired. If turns are removed here, it will be necessary to resonate with the loading coil, of course.

It should be emphasized that the figures shown in the table are calculated and therefore only approximate. They may be altered considerably depending on the type of car on which the antenna is mounted, and the loss resistances involved.

L Network

An *L* network may also be used to match a coax line to the whip antenna, as shown in Fig. 15-13. Values of X_L and X_C for matching a 50-ohm line to resistance loads may be determined from the curves of Fig. 15-12. For example, if it is desired to match a 50-ohm line to a 20-ohm resistance, X_C will be about 46 ohms, and X_L about 25 ohms.

Actually, a separate matching coil, L_M, need not be used, since increasing the inductive resistance of the loading coil by the equivalent amount, as indicated

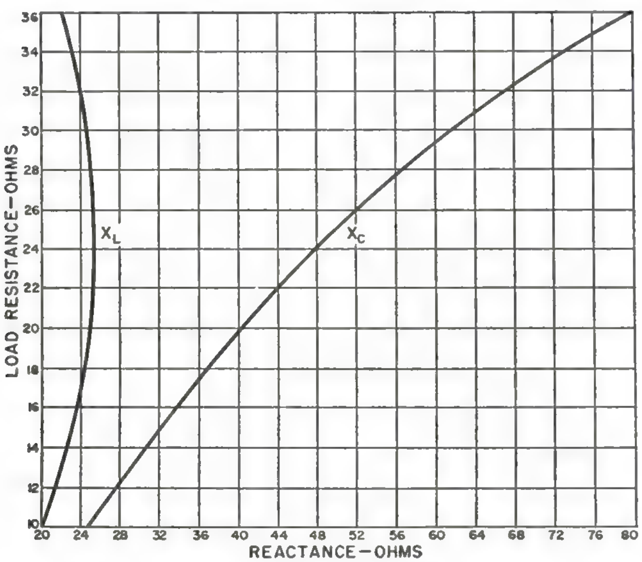


Fig. 15-12—Curves showing inductive and capacitive reactances required to match a 50-ohm coax line to a variety of antenna resistances.

at the right in Fig. 15-13, has exactly the same effect.

Since the feed-point resistance is seldom known with accuracy, this system is also best adjusted as described for the shunt-matching system above, preferably using an s.w.r. indicator.

TUNABLE WHIP ANTENNA SYSTEMS

At the lower frequencies, where the resistance of the circuit is low compared with the coil reactance, the antenna will represent a very high-Q circuit. This makes it necessary to retune the antenna for relatively small changes in frequency. Many methods have been devised for tuning the whip antenna system over a band, or from one band to another. Some use a series of plug-in coils, while others have a single coil with means for shorting out turns. In some cases, a small variable capacitor is used.

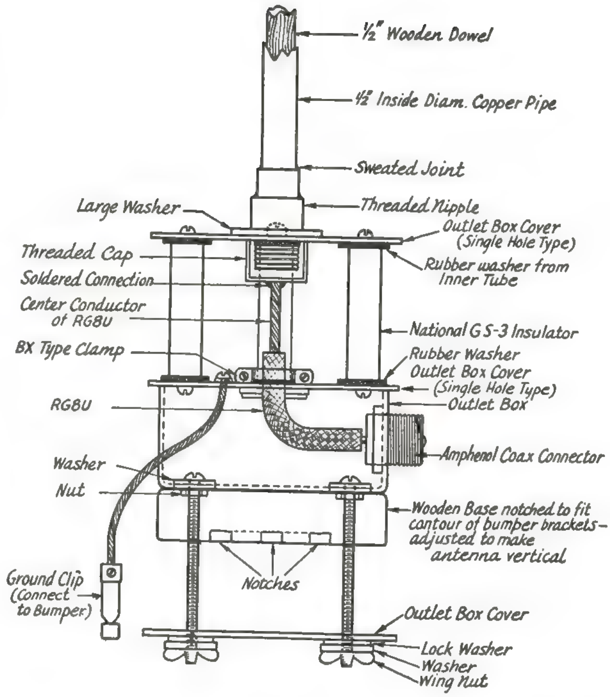


Fig. 15-14—Sketch showing the base mounting for the slider-tuned mobile antenna.

Slider-Tuned Whip

Figs. 15-14, 15-15, 15-16, and 15-17 illustrate a whip antenna with a homemade loading coil that is tunable over several bands by means of an adjustable slider that shorts out as much of the coil as necessary to tune to the desired frequency.

Most of the parts required in constructing the antenna are easily-obtained standard radio, electrical or plumbing items. In most cases it will be unnecessary to drill holes in the car. The mounting assembly is shown in the sketch of Fig. 15-14 and also in the photograph of Fig. 15-15. The mounting base is made principally of a standard four-way conduit junction box having a top cover with a single central knockout. A piece of ¾-inch wood is cut to match the bottom of the box and the bottom side of the wood is notched out to match the bumper support. A pair of long bolts (they usually come with the box) is inserted in the holes in the bottom lip of the box and clamped fast with nuts. The wood block is drilled to match the bolts and counter-bored for the nuts. The bottom metal cover of the box is used as a clamp for the bumper bracket. Wing nuts with lock washers are used to permit quick removal of the antenna assembly when desired.

One of the knockouts in the side of the box is drilled to take a coaxial-cable connector. The knockout hole in the top cover is fitted with a BX-clamp insert. Then a short section of RG-8/U cable is run between the connector and the clamp insert, leaving an extension of a couple of inches. A second outlet-box cover, similar to the other, is used as the base of the antenna. It is insulated from the box on three heavy ceramic pillars (National GS3), using a rubber washer at each end of each pillar.

The bottom section of the radiator is a piece of copper pipe ½-inch inside diameter and 36 inches long. At each end a brass nipple with a threaded neck is soldered. A piece of ½-inch wood dowel, filling the entire length of the pipe, is inserted to add strength and take out some of the whip. Then a large flat washer is slipped over the threaded neck of the nipple, the neck

Fig. 15-15—The mounting for the slider-tuned antenna is made up of standard parts and fittings as described in the text.

is inserted in the hole in the cover plate and fastened securely with a threaded pipe cap. The center conductor of the RG-8/U cable is soldered to the cap. The box is fitted with a length of flexible wire terminating in a clip that can be attached to the bumper for a ground connection. A coat of matching paint on the pipe makes it less noticeable on the car.

Figs. 15-16 and 15-17 show the details of the coil and slider mechanism. Although the coil should be wound on the best material available, preferably polystyrene, a wood dowel properly treated works very satisfactorily. It is 1⅜ inches in diameter and 12 inches long. The ends are turned down to make a snug fit into heavy surplus plug shells of the type that come fitted with cable clamps. The bottom end of the dowel is counterbored to take the threaded nipple at the top of the bottom antenna section and also a flat nut that fits the nipple threads. The top end of the form is drilled out to take the bottom end of the top section of the antenna which is a three-piece collapsible whip with a total length of 8 feet. The form is then boiled in paraffin for about a half hour, or until the bubbling ceases.

When completely cool and hard, the form is wound with 144 turns of No. 18 enameled wire, close-spaced, centering the winding on the form. The winding is 6¼ inches long. After winding, the entire coil is repeatedly dipped in paraffin to waterproof it. Then a path for the slider is sanded clean along the length of the coil.

The slider rods are of brass or copper, ¼-inch diameter. They are spaced by a short piece of ¼-inch-square rod at each end, the slider rods

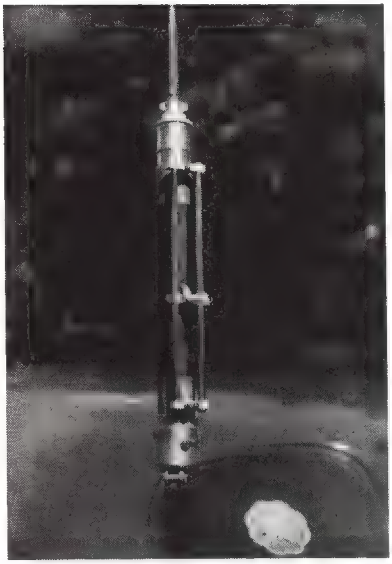


Fig. 15-17—Close-up of the slider-tuned loading coil.

being drilled and tapped for the assembly screws. The slider assembly is fastened to the coil form with wood screws and suitable spacers. The slider can be of odds and ends of copper and phosphor bronze found in the junk box. No tendency for the slider to shift with car vibration was noticed. The slider rods and contact and also the path along the coil should be touched up with sandpaper occasionally to remove oxidation.

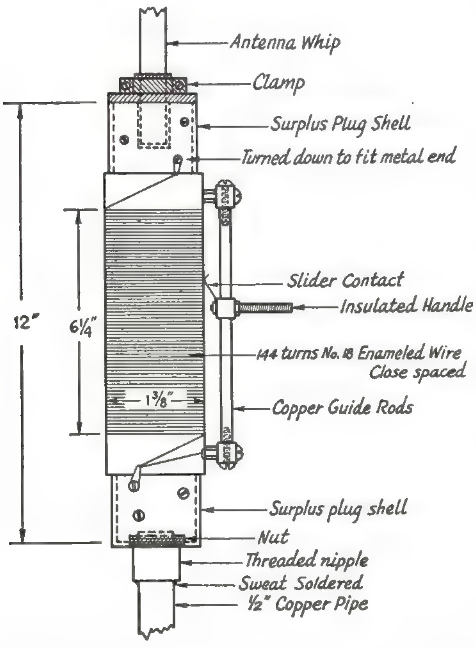


Fig. 15-16—Details of the slider loading-coil construction.

After the slider assembly has been mounted on the form, the cable shield at the bottom (minus the cable clamp) should be fastened to the top end of the copper pipe, screwing the nut down firmly in the bottom of the shell. It may be necessary to ream out the hole in the shell before it will slip over the nipple. Then insert the bottom end of the coil form into the shell and make it fast with several wood screws. Connect the bottom end of the coil winding and also the slider rods to the shell. Now slip the top shell on over the form and fasten it in the same manner, connecting the top of the winding to it. The top end of the slider assembly is left free, of course. The upper section of the radiator is held in place by the cable clamp on the top shell and can be removed as desired by loosening the clamp screws. The base section can be easily dismounted by loosening the wing nuts, thus leaving no sign of a mobile installation.

The antenna can be tuned up for use with one, two or three top sections attached to the loading coil, depending upon road-clearance conditions likely to be met. Naturally, the longer lengths give better results. The whole system can easily be brought to resonance by merely moving the slider to the correct and *exact* turn when the frequency is adjusted anywhere from 3500 to 4000 kc. or the length of the antenna above the load-

ing coil changed from three sections to one section (or any intermediate length). This, of course, applies to 20 and 40 meters as well.

Capacitive-Hat Tuning

Another simple and efficient method of tuning the antenna over a band is shown in the sketches of Figs. 15-18, 15-19, and 15-20. In this case, an old transmitting plug-in coil is used as the loading coil. A length of large-diameter polystyrene rod is drilled and tapped to fit between the upper and lower sections of a standard whip antenna. The assembly also serves to clamp a pair of metal brackets on each side of the polystyrene block, that serve both as support and connections to the loading-coil jack bar.

A ⅛-inch steel rod, about 15 inches long, is brazed to each of two large-diameter washers with holes to pass the threaded end of the upper section. The rods form a loading capacitance that increases as the upper rod is pivoted away from the lower one, the latter being stationary. Enough variation in tuning can be obtained in this manner to cover the 80-meter band. Fig. 15-18 shows the top washer slightly smaller than the lower one to facilitate marking a frequency scale on the stationary washer, after the upper washer has been marked with an index line. It is a good idea to hammer the washers into slightly cup shape to provide a little spring friction. The washers should be placed with the cups opposing.

After the movable rod has been set, it is clamped in position by tightening up on the upper antenna section. The plug-in mounting provides a convenient means of changing loading coils.

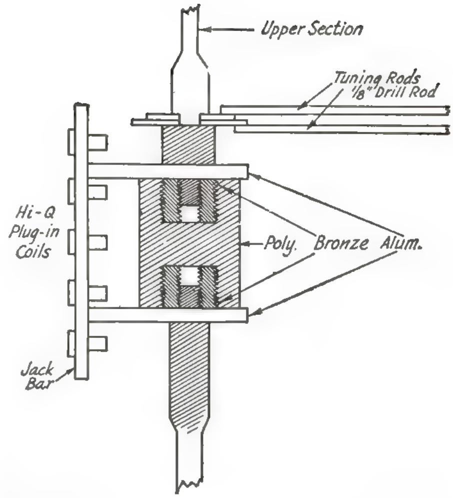


Fig. 15-19—Construction details of the mounting for the rods and plug-in coil.

If standard loading coils are used, turns should be removed until the antenna resonates at the high-frequency end of the band with the rods together. Start out with the rods perhaps 18 or 20 inches long, and prune them, bit by bit, until only the desired frequency range is covered. Most of the frequency change will take place with the rods relatively close together.

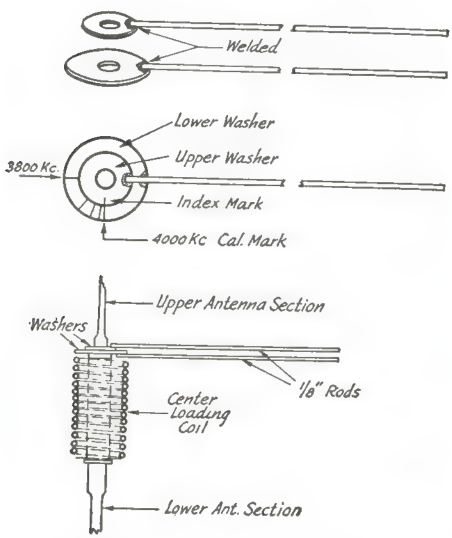


Fig. 15-18—Details of rod construction. Dimensions can be varied to suit the whip diameter and the builder's convenience. Adjustment of rod lengths is described in the text.

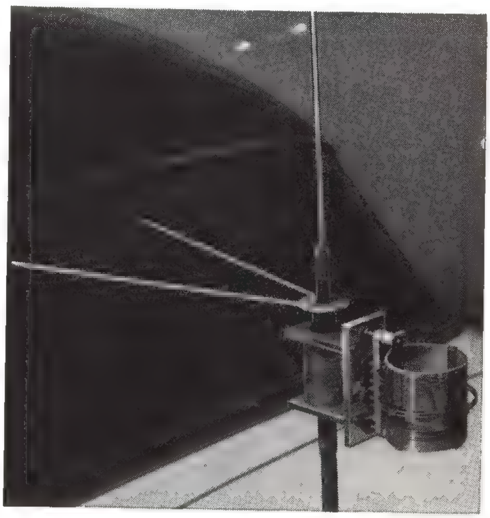


Fig. 15-20—W8AUN's adjustable capacity hat for tuning the whip antenna over the 75-meter band. The coil may consist of 45 turns No. 14, 3 inches in diameter, 10 t.p.i. Spreading the rods apart increases the capacitance. This simple top loader has sufficient capacitance to permit the use of approximately the same loading-coil inductance at the center of the antenna as would normally be required for base loading.

Remote Resonating with Motors

An interesting remote-control resonating system used by K6DY and W6WOY for their mobile antennas is shown in Figs. 15-21, 15-22 and 15-23. They make use of surplus twenty-four-volt d.c. motors driving a loading coil removed from a surplus ARC-5 transmitter.

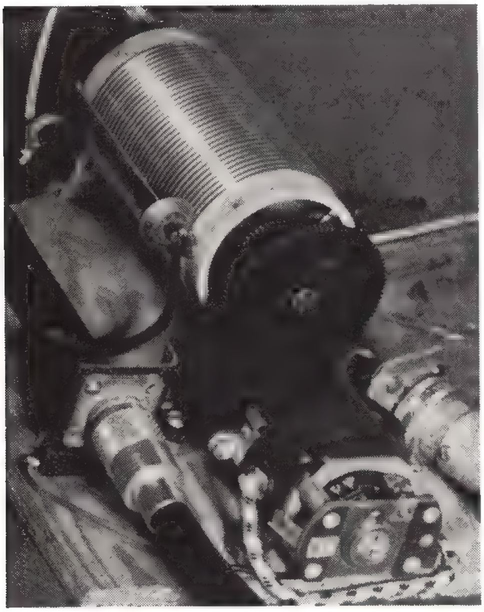


Fig. 15-21—The ARC-5 roller coil in this remotely-tuned antenna system is driven by a small pinion gear on the shaft of the surplus motor. The pinion fits the original fiber gear on the coil.

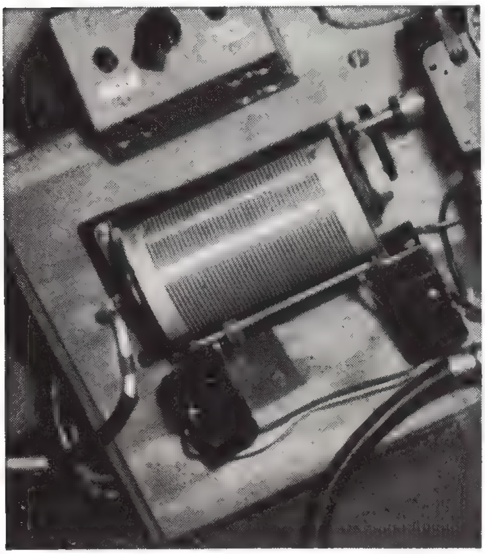


Fig. 15-22—The roller contact on K6DY's tuning coil actuates microswitches, placed at either end of the coil, to reverse the motor.

Many of the 24-volt surplus motor will run on 12 (or even 6) volts d.c. with sufficient torque to drive the coil. It may also be found that some of the motors are already equipped with gears that mesh perfectly with the fiber gear on the coil.

The control circuit used by W6WOY, shown in Fig. 15-23A, is a three-wire system (the car frame is the fourth wire) with a double-pole double-throw switch and a momentary (normally off) single-pole single-throw switch. S_2 is the motor-reversing switch. The motor runs so long as the push-button switch, S_1, is held closed.

K6DY introduces an additional refinement by using a latching relay, in conjunction with microswitches, so that the motor automatically reverses when the roller reaches the end of the coil. This circuit is shown in Fig. 15-23B. S_3 and S_5 operate the relay, K_1, which reverses the motor. S_4 is the motor on-off switch. When the tuning-coil roller reaches one end or the other of the coil, it closes S_6 or S_7, as the case may be, operating the relay and reversing the motor.

The procedure in setting up the system is to prune the center loading coil to resonate the antenna on the highest frequency used without the base loading coil. Then the base loading coil is used to resonate at lower frequencies. W6WOY throws S_2 (Fig. 15-23A) to "up" or "down," to increase or decrease frequency, and then controls the motor by means of S_1. K6DY momentarily closes S_3 or S_5 (Fig. 15-23B) to

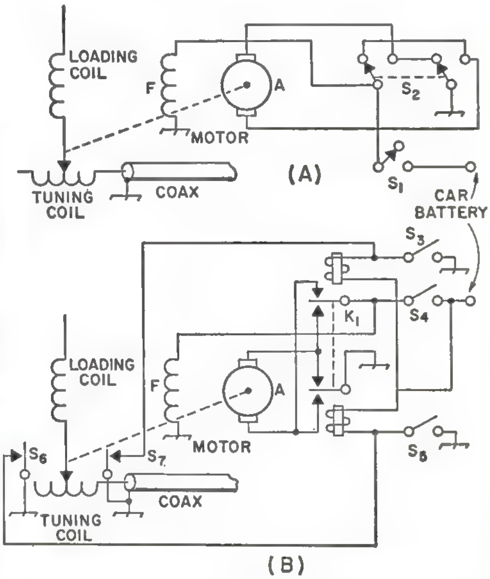


Fig. 15-23—Circuits of the remote mobile-whip tuning systems used by K6DY and W6WOY.

K_1—D.p.d.t. latching relay, 6- or 12-volt coil as required.

S_1, S_3, S_4, S_5—Momentary-contact, s.p.s.t., normally open.

S_2—D.p.d.t. toggle.

S_6, S_7—S.p.s.t. momentary-contact, normally open.

close the latching relay for changing frequency up or down, and then controls the motor with S_4. By using an additional latching relay, K6DY has pilot lights on the control panel to show in which direction the motor is running.

Using this system, it is possible to shift frequency while in motion without loss of signal strength. The broadcast antenna is used with a wavemeter to indicate resonance.

Another system of remote tuning is shown in Figs. 15-24 through 15-27. This system makes use of a salvaged electrically operated car broadcast antenna, and is applied to a Band Spanner antenna. The broadcast antennas are available in 6- and 12-volt models. These antennas are made so that the three sections telescope into the base. The motor drives a flexible nylon rod that is attached to the bottom of the top section. First remove this top section so that the nylon rod will push through the top of the middle section when power is applied to the motor lead that drives the antenna up. This operation can easily be done by using a sharp-edged file and cutting the top of the middle section just below the indented bead. After the nylon rod has been driven up a few inches, remove the top section from the brass coupling that attaches the nylon to the top section. This will leave about 3/16 inch of the brass coupling that can be tapped for 8-32 thread. Use a bottoming tap in order to get as much thread as possible.

The Band Spanner antenna is modified as follows: unscrew the top cap from the coil and pull the whole rod and slider assembly completely out of the coil tube. Place the rod in a vise and unscrew the brass nut to remove the circular contact assembly. When this has been done, remove the ball from the top of the rod (this is a compression fit), slide the cap assembly from the rod, replace the ball, thread the brass stud at the end of the rod from which the contact was removed with a 3/8 × 24 thread, and screw on a female coupling. This rod now becomes the top section of the antenna and is attached to what was originally the bottom part of the coil. The coil section has been inverted in order that the threaded end now at the bottom may be attached to the drive motor.

To complete the next step, a friend with a lathe is a big help. It will be necessary to make an insulated coupling in order to connect the two units together. Details of this coupling are shown in Fig. 15-27.

After the four separate parts of the antenna have been completed and a suitable location on the car trunk deck or cowl has been decided on, a rubber base and sleeve must be made so that the antenna, when mounted, will sit vertically on the car body for good appearance. When the proper size hole has been punched in the car deck and the complete antenna assembly installed, connect the motor leads to the switch on the dash as described in Fig. 15-25.

The coax connection to the antenna is made to the original electric antenna connector by making an adapter. This can easily be done by cutting the plug from the lead-in wire which came with the electric antenna and connecting it to a Type 83-1SP (PL-259) coax connector (see Fig. 15-24).

To relieve the strain on the coil section of the antenna, a small spring (Antenna Specialists Type M-25) is mounted on top of the coil. This does a good job of absorbing shock when passing under low tree branches and similar obstructions. A Master Mobile easy-off connector

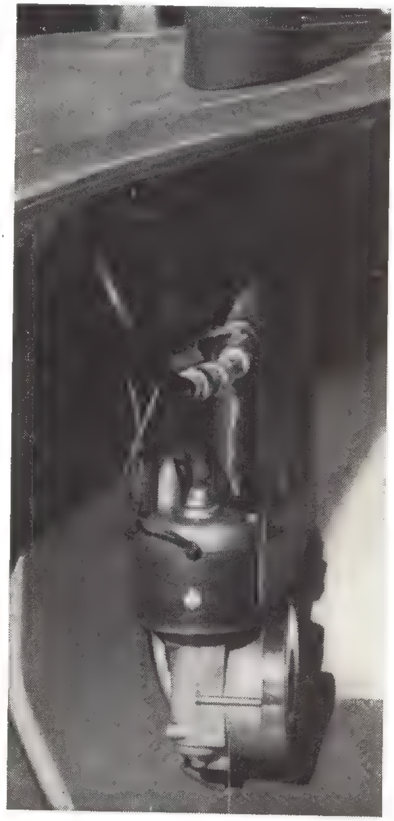


Fig. 15-24—A remote mobile antenna tuning system using the driving mechanism from a salvaged car broadcast-receiver installation (W1NI). The adaptation is applied to a Bandspanner antenna. This view shows the modified cable connector.

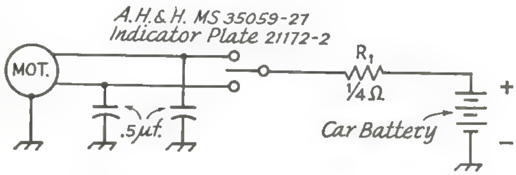


Fig. 15-25—Electrical control connections, including hash filter capacitors and series resistor, R_1, for adjusting motor speed.

is used for quick disassembly of the whip when entering garages. To prevent possible damage to the coil section when the car is passing through an automatic car-washing machine, this section can be quickly removed by running the drive motor to its lowest position, unscrewing the coil from the insulated coupling, and in its place screwing on a metal or plastic cap to prevent water from getting inside the coupling. This operation takes about two minutes and can be done while the car is being vacuumed before entering the washing machine.

The top section of the antenna can be made lighter, thereby reducing strain on the base section by substituting an Antenna Specialist type 19A328-1 whip for the original. With this substitution, a type 19A434-1 spring is used. When the whole assembly has been mounted, run two control wires to the front of the car and mount a single-pole double-throw momentary-contact switch on the dash and connect as shown in Fig. 15-25. The two leads should be bypassed with 0.5-μf. capacitors to eliminate hash in the receiver, and a ¼-ohm or larger resistor can be connected in the motor hot lead to reduce the speed of the slider, if necessary. The size of this resistor depends on the voltage and current of the motor used. The 12-volt models usually run from 5 to 8 amps.

The momentary switch will practically control antenna-coil tuning turn by turn, as desired, by a slight touch of the finger. An indicator plate is used on the switch. This plate reads HI in the up position and LOW in the down position, which corresponds to the inductance position of the slider on the coil. When the switch is held down the push rod moves to its lowest position,

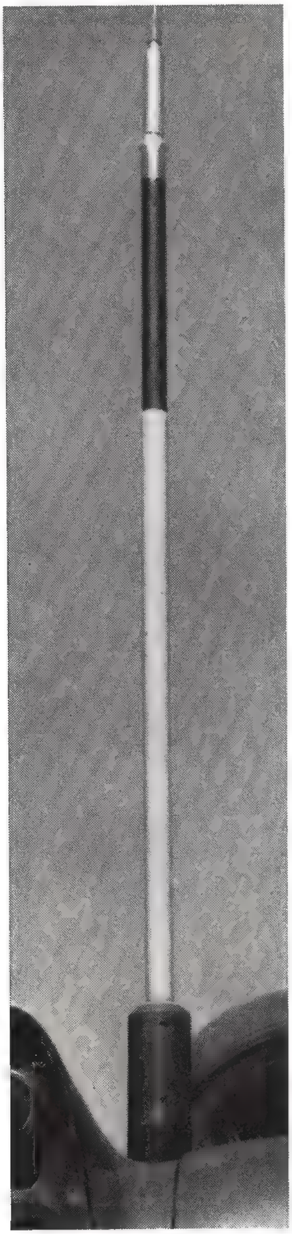


Fig. 15-26—Above-deck, view of the remotely-driven mobile antenna. The spring mentioned in the text is at the top in this view. The whip section, not visible, extends above. Details of the base mounting are shown in Fig. 15-27.

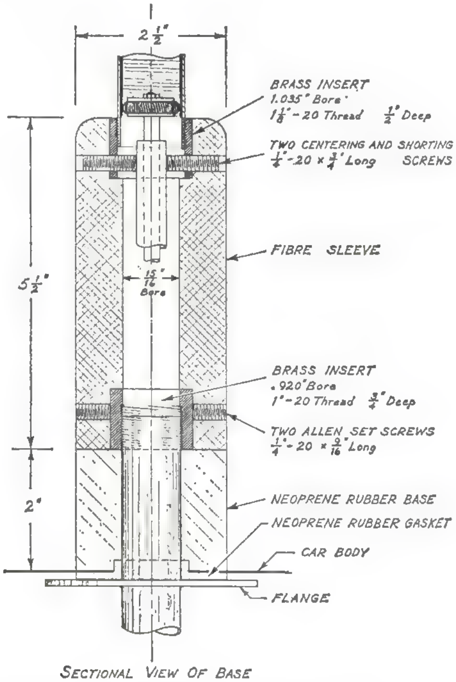


Fig. 15-27—Rear-deck mounting for the Band Spanner antenna and tuning drive. The fiber sleeve is made from a length of 2½-inch diameter linen-base phenolic rod, usually available from plastics suppliers (consult the Yellow Pages of the local telephone directory).

which is the maximum number of turns on the coil. Stops are unnecessary when the end of travel is reached because the nylon rod is driven by a slip clutch arrangement.

A Multiband Antenna System

The engineering staff of the E. F. Johnson Co. is responsible for the design of a loading network that permits multiband operation of a whip antenna on more than one band without changing coils, or retuning the antenna.

In Fig. 15-28A, which shows a dual frequency antenna with the lower of the two frequency bands designated F_2 and the higher F_1, the conductor length, X plus Y, is so chosen as to be resonant at F_1. The values of L_2 and C_1 are such that series resonance at F_1 occurs between terminals A and B. Since a series-resonant circuit offers zero impedance at the resonant frequency, terminals A and B are electrically short-circuited, thus still leaving the conductor, X plus Y, resonant at F_1.

Because there is zero impedance across terminals A and B at F_1, circuit elements L_1 and L_3 may be placed across terminals A and B without effect on the antenna behavior at F_1. Circuit element L_3 is of a value which, in conjunction with L_2 and C_1, forms a parallel-resonant circuit across terminals A and B at F_2. A parallel-resonant circuit presents infinite impedance across its terminals so that the combination of L_2, C_1 and L_3 is effectively not connected across terminals A and B at F_2.

The magnitude of L_1 is selected so that in conjunction with conductor X plus Y the system is resonant at F_2. The parallel combination of L_1, L_2C_1 and L_3, in itself is not resonant at F_2.

In practical application, either C_1 or L_2 is made variable to provide adjustment at F_1. L_1 and L_3 are combined in their parallel equivalent and the resultant inductor made variable to provide adjustment at F_2, as in Fig. 15-28B.

Conductors X and Y, for example, may be lower and upper portions of a center-loaded antenna with L_2C_1 series resonant at 10 meters and L_1 having a value which provides over-all resonance at 20 meters. The adjustments are made

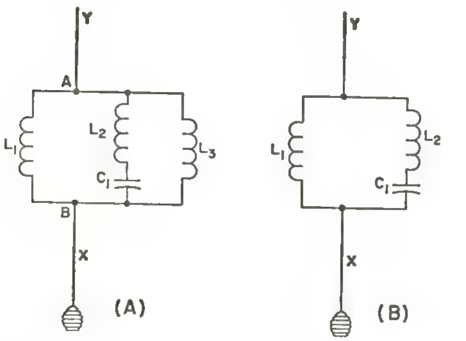


Fig. 15-28—Development of the two-band automatic switching network.

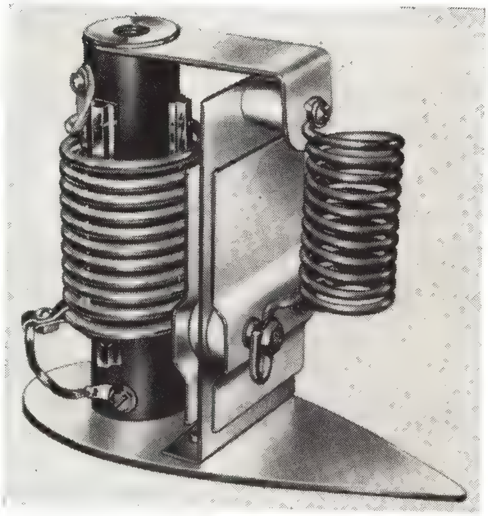


Fig. 15-29—A commercial version of the two-band networks.

by shorting L_2C_1, grid-dipping L_2C_1 at the center of the 10-meter band, removing the short and checking over-all antenna resonance on 10 meters.

If the over-all system is not resonant the top of section Y should be trimmed until resonance is obtained. The inductance of L_1 should then be adjusted so that the over-all system is resonant on 20 meters. The system is broad enough to cover the whole of the 10- and 20-meter bands without readjustment.

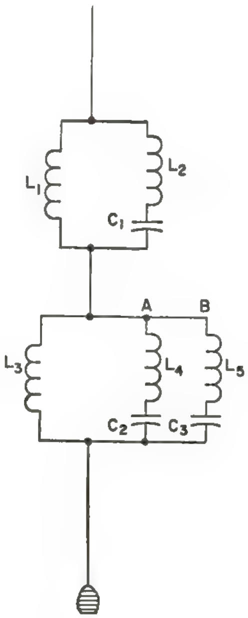


Fig. 15-30—Three-band automatic switching network.

Approximate values for 10- and 20-meter operation are 2.21 microhenrys for L_1, 0.85 microhenry for L_2 and 36 pf. for C_1. The photograph of Fig. 15-29 shows a prodcction version of the network in which L_1 and L_2 are made variable, so that the inductances in each branch may be adjusted for proper operation.

The antenna can be three-banded by the addition of another network as shown in Fig. 15-30. In this case the frequencies F_1, F_2 and F_3 are in decreasing order. L_2C_1 is series resonant at F_1, as is also L_4C_2. L_5C_3 is series resonant at F_2. The network $L_1L_2C_1$ is the same as shown in Fig. 15-28B, and is tuned to F_1 and F_2 as described above. L_3 which is electrically shorted at F_2 and F_1 by the series circuits connected at A and B, is of a value which will resonate the entire system at F_3.

V.H.F. Mobile Antennas

Simpler Types

The most convenient type of antenna for mobile v.h.f. work is the quarter-wave vertical radiator, fed with 50-ohm coaxial line. The antenna may be a telescoping whip if it is to be used for more than one band, or, in the case of 144 Mc. and higher bands, it may be simply a small rod cut to length. It may be mounted in any of several positions on the car, although the preferred spot is on the car top. This ideal may be possible only for the higher bands, however. Where the whip type of antenna is mounted in any position below the top of the car, marked directional characteristics are always in evidence; thus it is nearly always desirable to use the same antenna for transmission and reception.

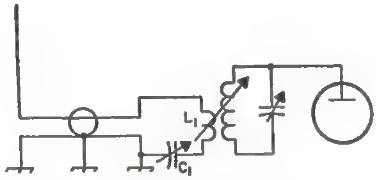


Fig. 15-31—Method of feeding quarter-wave mobile whip antennas with coaxial line. C_1 should have a maximum capacitance of 75 pf. for 28- and 50-Mc. work. L_1 is an adjustable link.

When coaxial feed is used in the mobile installation the coupling circuit should be arranged as shown in Fig. 15-31. The coupling should be adjustable, the optimum setting of C_1 being that which allows the loosest coupling. Inclusion of such a variable tuned coupling arrangement will make it possible to load the antenna at various lengths, a desirable feature if multiband operation is contemplated.

Two types of removable no-hole-mounting 144-Mc. whips are shown in Figs. 15-32 and 15-33. In Fig. 15-32, a coaxial fitting attached to a small strip of aluminum, bent so as to slip into the crack between the inside beading and the main frame of the car window. The weather-stripping in the car door flattens out around the aluminum strip and the coaxial line when the door is closed. Pressure against the mounting plate holds it in position, and the plate can be bent slightly after it is in place, to align it exactly vertical. By leaving a small amount of slack in the coax, the mount may be left in place as the door is opened and closed.

Another type of mounting, not so readily detachable, but one that permits mounting at the more favorable spot—the center of the car roof—is shown in Fig. 15-33. It was suggested by W6RLB.

It is made from the top of a tin can, a sheet of flashing copper and two coaxial fittings. One fitting is mounted in the top of the can, another in the side, and the two inner conductors are connected together inside the assembly. The can is cut about an inch high with three or four tabs extending a half inch or so for soldering to the copper sheet. The sheet is then fastened to the car top with plastic tape.

No direct electrical contact is needed between the plate and the car top, as the capacitance between the two simulates an actual ground connection. As a protection against scratching the car top, the bottom surface of the plate may be covered with a thin plastic sheet. If a lightweight whip is used, a strip of ⅜-inch tape around the edge of a 4- by 6-inch plate will hold the assembly firmly in place for months.

A collinear 144-Mc. antenna, suggested by W2ALR, is sketched in Fig. 15-34. The usual quarter-wave rod is made of stiff material so that it will support the phasing section and half-wave radiator above it. The phasing section and radiator are made and a detachable unit that can be screwed onto the top of the rod when the added coverage they provide is needed. The phasing

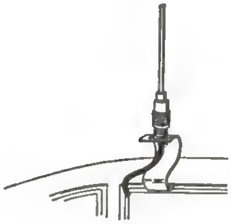


Fig. 15-32

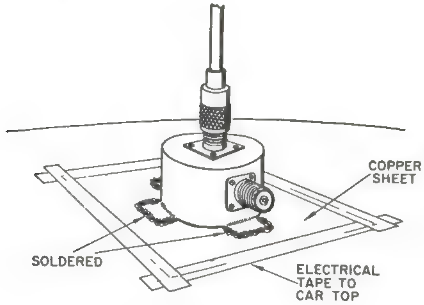


Fig. 15-33

line can be formed into a circle to cut down its over-all dimensions.

HORIZONTAL POLARIZATION

Experience has shown that a horizontally-polarized antenna has a considerable advantage over the vertical whip under most conditions of mobile operation. This is particularly true when horizontal polarization is used at both ends of a line-of-sight circuit, or a longer circuit over reasonably flat terrain. Additional advantage is gained, especially on 6 meters, from a marked reduction in ignition noise from neighboring cars as well as from the station car.

The Halo

One type of horizontally polarized antenna, called the halo antenna, is shown in Fig. 15-35. It is simply a folded dipole bent into a circle, with the ends capacitively loaded to reduce the over-all length.

For 50 Mc., the two rings are 20 inches in diameter. The upper conductor is of ⅞-inch tubing, while the lower conductor (feed-point side) is of ⅜-inch tubing. The two rings are separated 2½ inches, center to center. Details of the capacitor assembly, the dipole, and the mounting arrangement are shown in Fig. 15-36.

The spacing between the plates of the capacitor should be adjusted to resonate the system at the operating frequency. The spacing may be adjusted by adding or removing washers, as indicated in Fig. 15-36. The dipole dimensions given result in a feed impedance of 58 ohms, providing a good match for coax cable.

A similar antenna for 2-meter work is shown in Fig. 15-37. It consists of a folded dipole of No. 14 Copperweld wire, 39 inches long, bent into a circle about 12 inches in diameter. The conductors of the dipole are spaced ½-inch.

The outer ends of the dipole and the feed point are fastened to insulating blocks attached to a crossbar of aluminum tubing flattened at the ends. The halo can be supported in any convenient manner. In the photograph, the aluminum-tubing crossbar is flattened at the center and attached to the end of a vertical piece of similar tubing that slides down over the car's broadcast whip antenna.

The antenna may be fed with 300-ohm Twin-Lead or through a coaxial balun. The latter is to be preferred, since it permits grounding the coax

Fig. 15-34—Collinear mobile antenna used by W2ALR on 145 Mc. The phasing section and half-wave radiator are made in one unit that can be attached to the usual roof-top antenna. The phasing section is formed into a circle about 6 inches in diameter and fastened to the insulator between the two radiators.

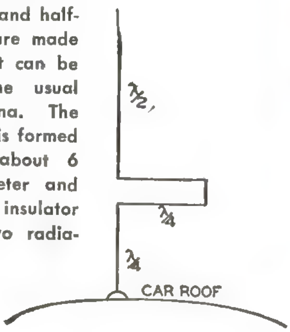


Fig. 15-35—The 50-Mc. halo used by W1MUX is mounted atop a pipe mast, the base of which is bolted to a metal angle plate welded to the rear bumper.

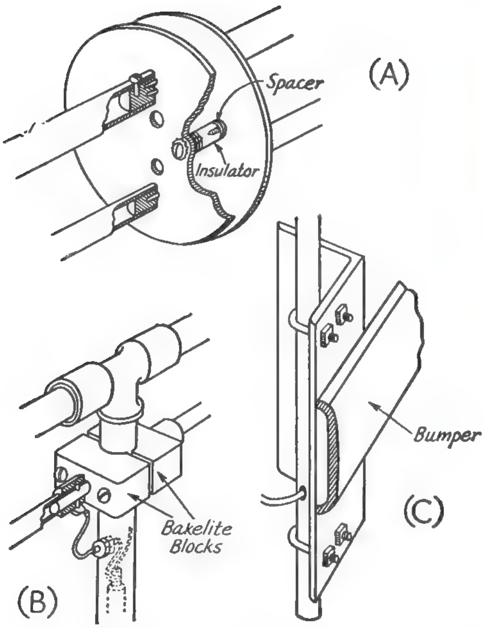


Fig. 15-36—Detail drawings of the various assemblies used in the halo installation. A is a cutaway view of the end-loading capacitor assembly. B shows the method of attaching the coaxial feedline and insulating the center of the fed section of the dipole. The method of attaching the mast to the rear bumper is shown at C. The angle plate is welded to the bumper assembly at three points.

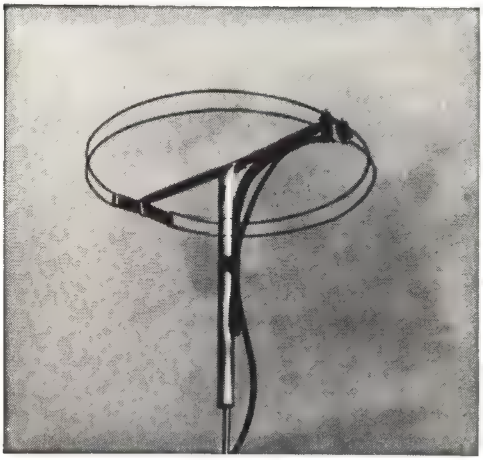


Fig. 15-37—A 2-meter halo by W3SST.

at any point along its run to the transmitter or receiver.

A Portable Beam for 2 and 6 Meters

Figs. 15-38 through 15-40 show the constructional details of a portable beam antenna that may be used with mobile 6- and 2-meter stations at fixed locations. Assembly or dismantling is a matter of only 10 minutes or so. The process of changing bands takes even less time.

The boom is made of two 3-ft. lengths of 1-inch aluminum tubing, end to end. A ¼-inch hole is drilled near the inner end of each section so that when the ends of the two sections are butted together, the holes will have proper spacing to fit the U bolt of a standard TV mast clamp equipped with wing nuts. Two 6-inch lengths of 1 × 1-inch aluminum angle are drilled with the same hole spacing. The angle pieces provide stiffening for the joint between the two boom sections. The clamp is illustrated in Figs. 15-39 and 15-40.

The elements are made of 5/16-inch dural tubing. Holes to clear the elements are drilled in the boom. The elements are clamped firmly to the boom as shown in Fig. 15-39. The clamp consists simply of two pieces of aluminum strip, drilled at one end to clear the element, and at the other to pass a machine screw. The pieces are bent roughly to conform to the boom periphery. Tightening the screw draws the sections of the clamp downward, clamping the element securely to the boom.

Starting at the rear (reflector) end of the boom, element holes should be drilled at ½, 21½, 33, 49, and 71½ inches from the end.

The table shows the lengths of the 2-meter elements. Both ends of the reflector (section No. 1), 1st director (section No. 3), and 3rd director (section No. 5) are threaded with a ¼-20 tap. The 2nd director (section No. 4) is tapped at one end. All tapped holes, except the one in the 2nd director, should be fitted with ¼-20 threaded studs of aluminum or brass as indicated by Fig. 15-39. After threading the studs in securely, they should be cut off so that they extend ½ inch from the ends of the elements.

Conversion to 6 Meters

The 5-element 2-meter array is converted to a 3-element antenna for 6 meters by threading tapped extensions onto each of the studs of the 2-meter reflector, 1st director (which now becomes the 6-meter driven element), and 3rd director. The lengths of these extensions are shown in the table. The driven element of the 2-meter beam is not used on 6 meters, but it may be left in its original position. The 2nd 2-meter director is unclamped from the boom and is used as one extension for the 6-meter director.

Matching Systems

Gamma match to RG-58/U or RG-59/U is used on both driven elements. The inner ends of the gamma sections (also of 5/16-inch dural tubing) are attached to the driven elements by means of ⅜-inch ceramic cone insulators, as shown in Fig. 15-40. Their lengths should be

Fig. 15-38—A 5-element 2-meter beam becomes a 3-element 6-meter beam in a matter of seconds. Four extensions all the same length screw onto the reflector and middle element. One of them is the second 2-meter director, which is slipped out of the boom for this purpose. Two shorter extensions screw onto the forward director. The coaxial cable is shifted from one gamma section to the other, and the 6-meter beam is ready for use.

about 5 inches for 144 Mc. and 13 inches for 50 Mc. The outer end of each gamma section is fitted with an adjustable shorting clamp.

The gamma series capacitors and coax receptacles are mounted on U-shaped aluminum brackets fastened permanently to the boom with sheet-metal screws. The rotor of the capacitor must be insulated from the bracket. The capacitors used are shaft-type (Hammarlund MAPC) trimmers, 50 pf. for 2 meters and 75 pf. for 6 meters. These capacitors have insulated mounting studs. The rotors are connected to the inner ends of the gamma sections by flexible copper straps fastened under the standoff insulator screw. This facilitates assembly and disassembly. The stators are connected permanently to the ungrounded terminals of the coax receptacle. A single coaxial cable is transferred from one driven element to the other in shifting bands.

Mast

The supporting mast is made up of three 5-ft. lengths of sectional TV mast. If 5-ft. lengths will not fit into the car trunk, they may be shortened at some sacrifice in antenna height, or a fourth section may be added. To keep the sections from turning individually in the wind, a notch is filed in the untapered bottom end of each section. Then a sheet-metal screw is threaded into the tapered end of the section below so as to engage in the notch.

A screwdriver driven into the ground may serve as a piot for the bottom end of the mast. In locations where this is not feasible, rocks piled loosely about the base will serve the same purpose. The mast can be braced against the side of the car by a bracket, fashioned from aluminum, that attaches to the door handle or window

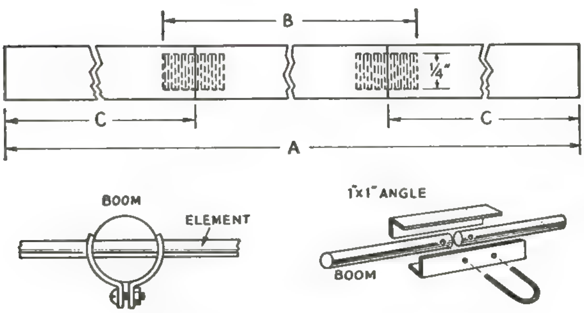


Fig. 15-39—Details of the portable beam antenna for 6 and 2 meters. In the upper drawing, dimension B is the basic 144-Mc. element length which includes the ½-inch threaded studs. C indicates the extensions added at each end of the 144-Mc. elements for 50-Mc. operation. Lengths for B and C are given in the accompanying table.

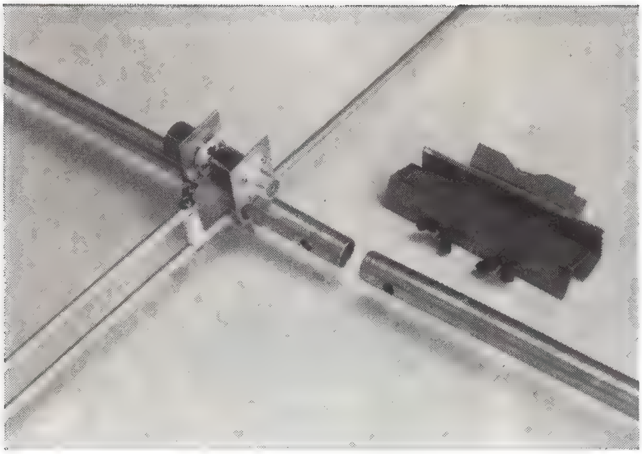


Fig. 15-40—Detail view of the 50-Mc. driven element, showing the tuning capacitor and gamma matching arm. The clamp that holds the two portions of the boom together, and to the vertical support, is seen at the right.

crevice. The section above the bottom section has a 4-inch ¼-20 bolt through it at a point where the bolt just rides on the top end of the bottom section. This makes a bearing surface inside the mast, and the portion of the bolt extending on the outside serves as a turning handle and direction indicator.

Adjustment

In adjusting the gamma match, an s.w.r. bridge should be used. The position of the shorting clamp and the setting of the series capacitor should be varied, first one and then the other, until the reflected power indication is made to go to zero. If this is done at some frequency near the middle of the operating range, the setting will usually be satisfactory over the major portion of the 2-meter band, and over better than one megacycle on 6 meters.

The Two-Meter "Big Wheel" Antenna

The 2-meter antenna shown in Fig. 15-41 is omnidirectional and horizontally polarized and has appreciable gain over other simple forms of antennas having similar characteristics, such as the halo and turnstile. In effect, it consists of three half waves in phase with end feed. Each element is a full wavelength, as shown in Fig. 15-42, bent to form a half-wave circumferential radiating segment with quarter-wave radial feeders. One feeder goes to the center common ground point, the other to the common high-potential point, fed by the center conductor of a coaxial line. The driving impedance is about 12 ohms. To match a 50-ohm line, a conventional stub-tuner arrangement is used. Element lengths are chosen so that the impedance is capacitive, and the circuit is then tuned to resonance with an inductive stub to give an input impedance of 50 ohms at the center frequency.

Constructional details are shown in Fig. 15-43. Elements are made of ⅜-inch o.d. corrosion-resistant aluminum tube (Alcoa type 6061-T6). The lengths are bent to the shape shown in Fig. 15-43. This sketch also shows the mounting brackets and other hardware. Wood dowels are used to plug the element ends to add strength and seal against moisture. Each element (A)

Dimensions of 2- and 6-Meter Portable Beam

Sec. No.	*Element*	*Length (In.)*
1	144-Mc. Reflector	40½ *
2	144-Mc. Dr. El.	38½
3	144-Mc. 1st Director	36½ *
4	144-Mc. 2nd Director	36¼ **
5	144-Mc. 3rd Director	35½ *
6–7	50-Mc. Ref. Ext.	37½
8–9	50-Mc. Dr. El. Ext.	37½
10	50-Mc. Director Ext.	36¼

* Length includes ½-inch stud at each end.
** Serves as 2nd extension section of 50-Mc. director.

Fig. 15-41—The 2-meter "Big Wheel" antenna for mobile of fixed-station use. (W1IJD and W1FVY)

runs from the grounded plate (B) to the triangular plate (C). The two plates are mounted, one above the other, at a spacing determined principally by available insulators. Ceramic stand off insulators 1 to 1½ inches long (such as the Johnson type 135-501) are suitable. The tuning stub (D) is bent around a ¾-inch radius, but this is not critical.

Two or more wheel antennas can be stacked to increase the gain. Stacking increases the directivity in the vertical plane only; horizontal polarization and the omnidirectional pattern are preserved.

The gain of a stacked array depends on the spacing between bays as well as the number of bays. A spacing of ⅝ wavelength is about optimum and will give a gain about equivalent to a 2-element Yagi, but in all directions. With this spacing, it is convenient to use full-wavelength phasing lines of coax, as shown in Fig. 15-44. At 146 Mc., ⅝ wavelength is approximately 50 inches, while a full-wavelength coaxial line is only about 53 inches because of the velocity factor of the line. In Fig. 15-44 a full wavelength of RG-11/U 75-ohm coaxial line is used as the phasing section. It is driven by 50-ohm RG-8/U transmission line at a point ¼ wavelength up from the bottom to achieve the proper impedance transformation. The two ends of the coax are out of phase, so one of the bays must be

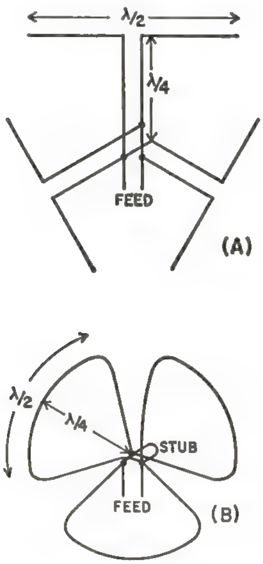


Fig. 15-42—The development of the "Big Wheel" is based on the arrangement of 3 half waves in phase shown at A.

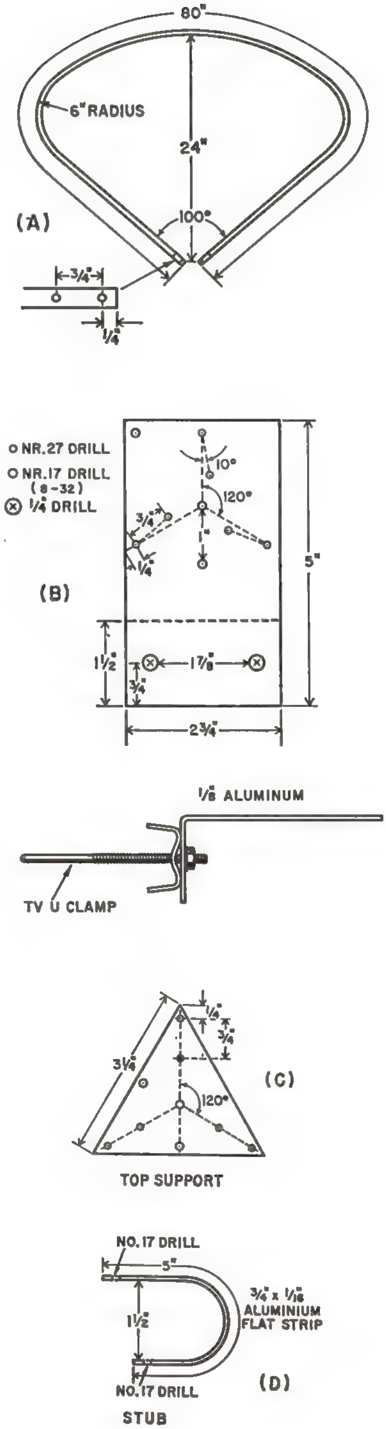


Fig. 15-43—Structural details of the "Big Wheel." One element is shown at A. The wood dowel, 2 inches long, is driven into each element end.

The grounded lower support is shown at B. It is bent down at right angles to permit mounting to a vertical pipe with a U clamp. The triangular top support is shown at C, and the tuning stub at D. Brass screws (¾-inch 6-32) are used in the assembly.

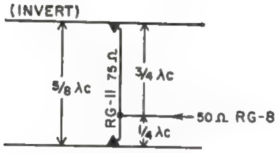


Fig. 15-44—Stacking method for two "Big Wheel" antennas. Because of the propagation factor of coaxial line, an electrical full wavelength of coax, λ_c, is approximately ⅝ wavelength long. This is the optimum stacking dimension for dipoles. By using a 75-ohm phasing line, the system may be fed at the point indicated with a 50-ohm transmission line. Note that one bay must be inverted to keep antenna currents in phase.

turned upside down to put the antenna currents in phase. Because of the coupling between the bays, the stub length must be increased to 6 inches total length. For a more nearly perfect circular pattern, the loops of the two bays should be staggered 60 degrees on the mast.

The gain may be increased still further by stacking four bays as shown in Fig. 15-45. In this arrangement, the top and bottom stubs are 6 inches long, while the two inner pair should be 7 inches long. The top and bottom bays should be inverted as indicated.

The Skew-Planar Wheel

The Skew Planar Wheel antenna is similar in principle and construction to the "Big Wheel," but four elements are used and the elements are twisted at an angle of 45 degrees to the horizontal, as shown in Figs. 15-46 and 15-47 to produce circularly polarized radiation. The input impedance is about 40 ohms, so a stub is necessary only when a perfect match is required.

The construction of the upper support for this antenna is shown in Fig. 15-48. The bottom support is the same as described for the "Big Wheel."

Because of the vertical radiation component, the Skewed Wheel must be mounted at the very top of the mast. For the same reason, this version cannot readily be stacked on a mast.

Element lengths for other frequencies may be determined in inches by dividing 11,800 by the desired center frequency in megacycles. The radius of the large bends is about 30 per cent of this length, and the radials about 25 per cent. The angle between the radials of each element is about 100 degrees.

Fig. 15-46—The Skew Planar Wheel mounted on the rear deck of W1FVY's car. (W1IJD and W1FVY)

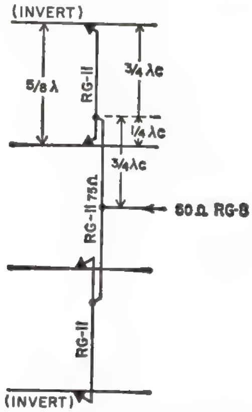


Fig. 15-45—Feed system for a four-bay "Big Wheel" array. The center bays are the same side up, while the two outer bays are inverted. Bays are approximately ⅝ wavelength apart physically, which permits the use of full-wave phasing sections between them. The feed points of each pair are then fed through two ¾-wave phasing sections, and a 50-ohm line at the mid-point sees an almost perfect match. The tuning stubs on the two inner bays are 7 inches long, while those on the outer bays are 6 inches long.

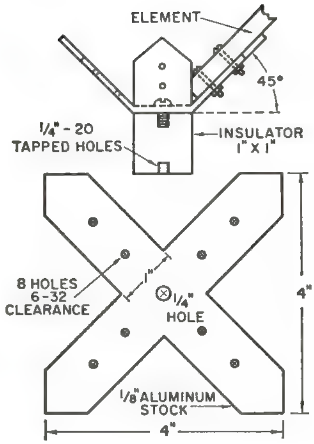


Fig. 15-47—(Left) close-up view of the mounting of the Skewed Wheel. Each element connects from the upper support to the lower, as in the original model of the antenna. Lower mounting plate is grounded to the support, and connected to the outer conductor of the coaxial feed line.

Fig. 15-48—(Right) construction of the Skewed Wheel is similar to the previously-described Big Wheel, except for the top support, and the use of four elements instead of three. A square plate of ⅛-inch aluminum is cut and drilled as shown, and the projections bent up to a 45-degree angle. Dimensions are not critical, so long as holes in the elements and the support match up.

Antennas for Transmitter Hunting

Hidden-transmitter hunts have become increasingly popular, and for successful participation in this activity a directional receiving antenna is essential. Such an antenna is usually in the form of a small loop tuned to resonance with a capacitor.

There are several factors that must be considered in the design of a direction-finding loop. The loop must be small compared with the wavelength. In a single-turn loop, the conductor should be less than 0.08 wavelength long. Thus at 29 Mc. the length should be less than 32 inches (diameter 10 inches).

To obtain most accurate bearings, the loop must be balanced electrostatically with respect to ground. Otherwise, the loop will exhibit two modes of operation. One is the mode of a true loop, while the other is that of an essentially nondirectional vertical antenna of small dimensions ("antenna effect"). The voltages introduced by the two modes are not in phase and may add or subtract, depending upon the direction from which the wave is coming.

The theoretical true loop pattern is illustrated in Fig. 15-49A. When the antenna effect is appreciable and the loop is tuned to resonance, the loop may exhibit little directivity, as shown at B. However, by detuning the loop so as to shift the phasing, a pattern similar to C may be obtained. Although this pattern is not symmetrical, it does exhibit a null. The null may, however, not be as sharp as that obtained with a loop that is well balanced, and may not be at exact right angles to the plane of the loop.

By suitable detuning, the undirectioal pattern of Fig. 15-49D may be approached. This adjustment is sometimes used in transmitter hunting to obtain a unidirectional bearing. In most cases, however, the loop is adjusted for the best null.

An electrostatic balance can be obtained by shielding the loop. If the loop is balanced, the two nulls should be 180 degrees apart. Some-

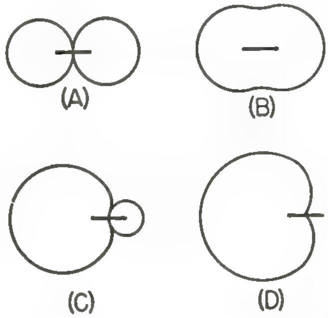


Fig. 15-49—Small-loop field patterns with varying amounts of "antenna" effect. The heavy lines show the plane of the loop.

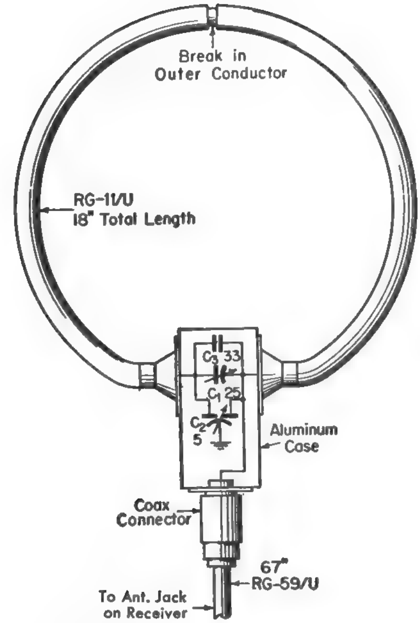


Fig. 15-50—Sketch showing constructional details of W9PYG's transmitter-hunt loop. The outer braid of the coax loop is broken at the center of the loop. The gap is covered with waterproof tape. The entire assembly is given a coat of plastic spray.

times the car broadcast antenna will interfere with accurate bearings. Disconnecting the broadcast antenna should eliminate this trouble.

LOOP CONSTRUCTION

Figs. 15-50 and 15-51 show the construction and mounting of a simple shielded 10-meter loop. The loop is made from an 18-inch length of RG-11/U secured to an aluminum box of any convenient size, with two coaxial cable hoods (Amphenol 83-1HP/U). The outer shield must be broken at the exact center. C_1 is a 25-pf. variable capacitor, and is connected in parallel with a 33-pf. mica padder capacitor, C_3. C_1 must be tuned to the desired frequency while the loop is connected to the receiver in the same way as it will be used on the hunt. C_2 is a small differential capacitor (Johnson 6MA11) used to provide electrical symmetry. The lead-in to the receiver is 67 inches of RG-59/U cable.

The loop can be mounted on the roof of the car with a rubber suction cup, although reasonably true bearings may be obtained through the windshield when the car is pointed in the direction of the hidden transmitter. More accurate bearings may be obtained with the loop held out the window and the signal coming toward that side of the car.

Another method of mounting the loop antenna so that it can be easily rotated by hand from the window is shown in Fig. 15-52.

Sensing Antenna

A sensing antenna can be added to the loop antenna so that a check can be made, as found desirable, on which of the two directions indicated by the loop is the correct one. In Fig. 15-50, a phono jack to take a short vertical antenna rod of a diameter to fit the jack can be mounted in the top of the aluminum box. The insulated terminal of the jack should be connected to the side of the tuning capacitors connected to the center conductor of the RG-59/U coax feed-line. The sense antenna can be plugged in as needed. Starting with a length of about four times the loop diameter, the length of the sensing antenna should be pruned until the pattern is similar to Fig. 15-49D.

D.F. Loop for 75 Meters

Figs. 15-53 and 15-54 show a d.f. loop suitable for the 3500-4000-kc. band using a construction technique that has had considerable application in low-frequency marine direction finders. The loop is a coil wound on a ferrite rod from a

Fig. 15-51—W9PYG's shielded 10-meter d.f. loop may be mounted on the roof of the car with a rubber suction cup.

broadcast-antenna "loopstick." Because it is possible to make a coil of high Q with the ferrite core, the sensitivity of such a loop is comparable with that of a conventional-type loop a foot or so in diameter. The output of the vertical-rod "sensing" antenna, when properly combined with that of the loop, gives the system the cardioid pattern shown in Fig. 15-49D.

To make the loop, remove the original winding on the ferrite core and wind a new coil as shown in Fig. 15-54. Other types of cores may be used than the one specified; use the largest core available and adjust the winding so that the circuit resonates in the 75-meter band within the range of C_1. The tuning range of the loop may be checked with a grid-dip meter.

The sense system consists of a 15-inch whip, an adjustable inductance that will resonate the whip as a quarter-wave antenna, and a potentiometer to control the output of the antenna. The switch S_1 is used to disconnect the sense antenna during the tune-up procedure.

The whip, the loop stick, the inductance L_1, the capacitor C_1, the potentiometer R_1, and the switch S_1 are all mounted on a 4 × 5 × 3-inch box chassis as shown in Fig. 15-55. The loopstick is mounted and protected by means of a piece of ½-inch thick laminated plastic and a length of fiber tubing which fits over the entire loop stick. A section of ½-inch electrical conduit is attached

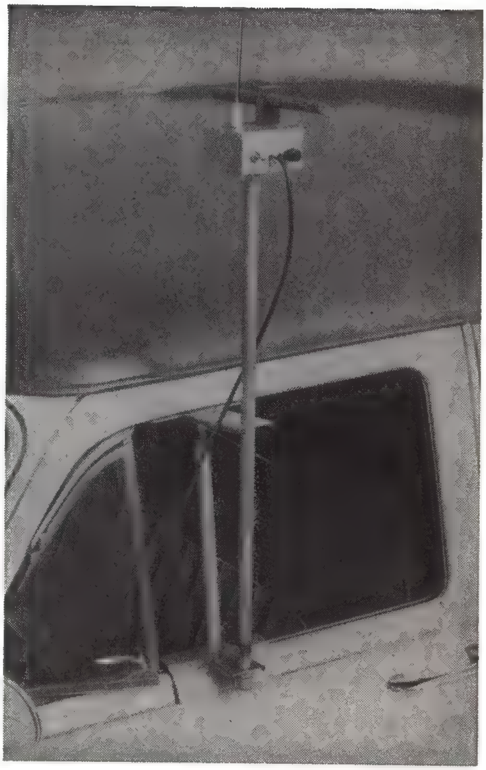


Fig. 15-53—Unidirectional 75-meter d.f. using ferrite-core loop with sense antenna. Adjustable components of the circuit are mounted in the aluminum chassis supported by a short length of tubing (W6PZV).

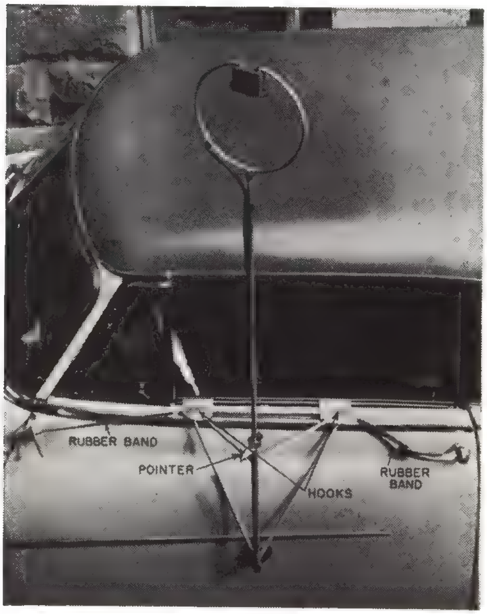


Fig. 15-52—A 28-Mc. direction-finding loop mounted on a car. The mounting is a triangular framework of tubing or rods with plates that hook over the window frame, and a rubber suction cup at the bottom. The loop mast revolves in a section of tubing. Large rubber bands to the external rear-view mirror and door handle help to hold the assembly in place (W7OTA).

to the bottom of the chassis box and this supports the d.f.

To produce an output having only one null there must be a 90-degree phase difference between the outputs of the loop and sense antenna and the signal strength from each must be the same. The phase shift is secured by tuning the sense antenna slightly off frequency by means of the slug in L_1. Since the sensitivity of the whip antenna is greater than that of the loop, its output is reduced by adjusting R_1.

To adjust the system, enlist the aid of a friend with a mobile transmitter and find a clear spot where the transmitter and d.f. receiver can be separated by several hundred feet. Use as little power as possible at the transmitter. (Remove your own transmitter antenna before trying to make any loop adjustments and remember to leave it off during transmitter hunts.) With the test transmitter operating on the proper frequency, disconnect the sense antenna with switch S_1 and peak the loop stick using C_1 while watching the S meter on the receiver. If no S meter is available one should be installed before the direction finder project is started. Once the loop stick is peaked, no further adjustment of C_1 will be necessary. Next, connect the sense antenna

and turn R_1 to minimum resistance. Then vary the adjustable slug of L_1 until a maximum reading of the S meter is again obtained. It may be necessary to turn the d.f. a bit during this adjustment to obtain a larger reading than with the loop stick alone. The last turn of the slug is quite critical and some hand capacitance effect may be noted.

Then turn the d.f. so that one side (not an end) of the loop stick is toward the test transmitter. Turn R_1 a complete revolution and if the proper side was chosen a definite null should be observed on the S meter for one particular position of R_1. If not, turn the d.f. 180 degrees and try again. This time leave R_1 at the setting which produces the minimum reading. Now adjust L_1 very slowly until the S-meter reading is reduced still further. Repeat this several times, first R_1

Fig. 15-54—Circuit of the 75-meter direction finder.

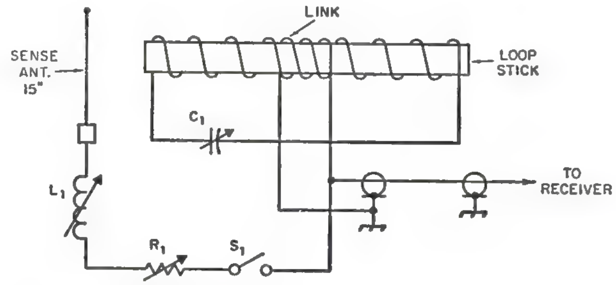
C_1—140 pf. variable (125-pf. ceramic trimmer in parallel with 15-pf. ceramic fixed.

L_1—Approx. 140 μh., adjustable (Miller No. 4512 or equivalent.-

R_1—1000-ohm carbon potentiometer.

S_1—S.p.s.t. toggle.

Loopstick—App. 15 μh. (Miller 705-A, with original winding removed and wound with 20 turns of No. 22 enam.) Link is two turns at center. Winding ends secured with Scotch electrical tape.



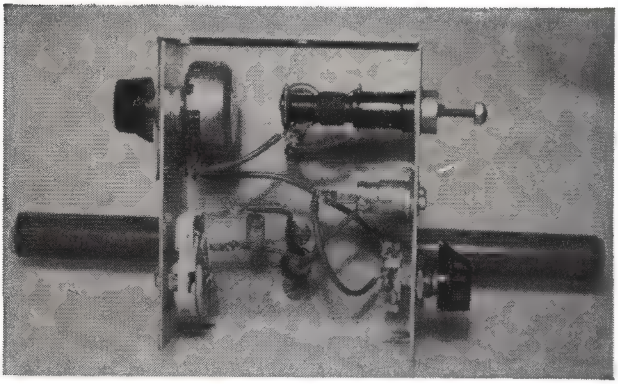


Fig. 15-55—Components of the 75-meter d.f. are mounted on the top and sides of a "Channel-lock" type box. In this view R_1 is on the left wall at the upper left and C_1 is at the lower left. L_1, S_1 and the output connector are on the right wall. The loop stick and whip mount on the outside.

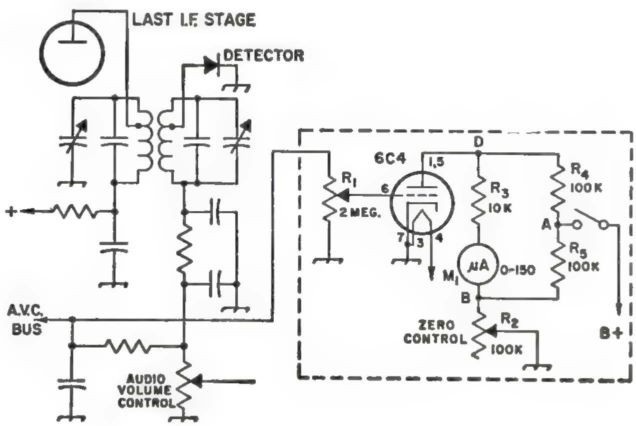


Fig. 15-66—Schematic diagram showing the circuit of a bridge-type signal-strength meter with sensitivity control, and how it is connected to the receiver a.v.c. bus.

and then L_1, until the best minimum is obtained.

Finally, as a check, have the test transmitter move around the d.f. and follow it by turning the d.f. If the tuning has been done properly the null will always be broadside to the loop stick. Make a note of the proper side of the d.f. for the null and the job is finished.

Signal-Strength Meters for D.F. Work

Difficulty is usually encountered in trying to tell the difference between maximum and minimum signal by ear as the hunter closes in, and a signal-strength meter becomes very desirable. An S-meter circuit of the amplifier-bridge type is shown in Fig. 15-56. This system operates by sampling the a.v.c. voltage, amplifying the voltage change, causing a change in the plate resistance of the 6C4 tube. This change in plate resistance upsets the balance of the bridge circuit, causing a difference of potential to exist between points *D* and *B*. The resulting current flow through the meter causes the needle to deflect. Potentiometer R_1 is a gain control and governs, to a certain extent, the amount of deflection of the meter. Potentiometer R_2 is the zero adjustment used to balance the bridge. As the signal of the hidden transmitter changes in intensity, both the gain and zero controls will need adjusting.

A 1-ma. meter will work quite well as a substitute for the 150-μa. unit, and almost any medium-μ triode may be substituted for the 6C4.

FIELD-STRENGTH INDICATORS

In adjusting a mobile transmitter and antenna for maximum output, a field-strength indicator is indispensable. Such an indicator and its circuit are shown in Figs. 15-57 and 15-58. Aside from its usefulness in making initial adjustments, the instrument can be mounted permanently in the car and used as an output indicator when changing transmitter frequency. It will also provide a continuous check on transmitter performance.

Nearly any type of crystal diode may be used, and the meter scale may be anything from 50 μa. to 2 ma. or more, since the position of the pick-up antenna in relation to the transmitting antenna can be changed to give the desired deflection in any event. The meter shown in Fig. 15-57 has a 1-ma. scale and is illuminated (Simpson Model 127). It is mounted in a standard aluminum meter box fitted with rubber feet.

L_1 is a "Vari-Loopstick" (a slug-tuned inductor used widely as a broadcast antenna). Once it is peaked for 75 meters, no further adjustment is necessary, since the sensitivity will be sufficient for the high-frequency bands without retuning.

The unit may be placed on top of the instrument panel or on the deck back of the rear seat. The length of the probe needed will depend

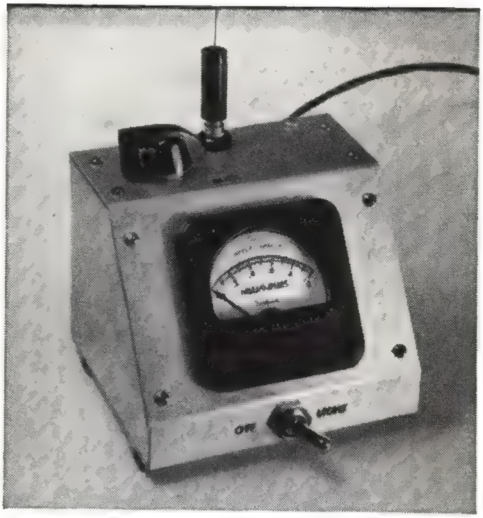


Fig. 15-57—The sensitivity control of this simple f.s. meter is mounted on top of the standard aluminum (Bud) meter case used as a housing. The pick-up probe is plugged into a phone-tip jack.

upon the sensitivity of the meter used, the transmitter power, and the distance of the probe from the transmitting antenna. In most cases, a length of 6 to 8 inches should be sufficient. The broadcast antenna, or even a fender guide if it is insulated from the car body, may also be used.

Combination Signal- and Field-Strength Indicator

An arrangement in which the functions of signal-strength meter and field-strength indicator are combined in a single unit is shown in Figs. 15-59 to 15-61, inc. With S_1 in the position shown in the diagram, the meter is connected in the S-meter circuit; the other switch position inserts the meter in the circuit of the field-strength indicator. R_1 is the sensitivity control for the S meter, and R_2 is a zero-adjust for the S meter, or a sensitivity control for the f.s. indicator. Power for the S-meter circuit (about 2 watts) may be taken from the car broadcast receiver. No power is required for the f.s. indicator.

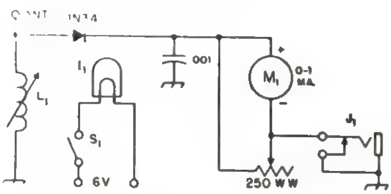


Fig. 15-58—Circuit of the simple f.s. meter. L_1 is a "Vari-Loopstick" (broadcast antenna type) adjusted to approximately 200 μh. J_1 is a closed-circuit phone jack for aural monitoring. S_1 is a s.p.s.t. toggle switch, and I_1 is a 6- or 12-volt dial lamp.

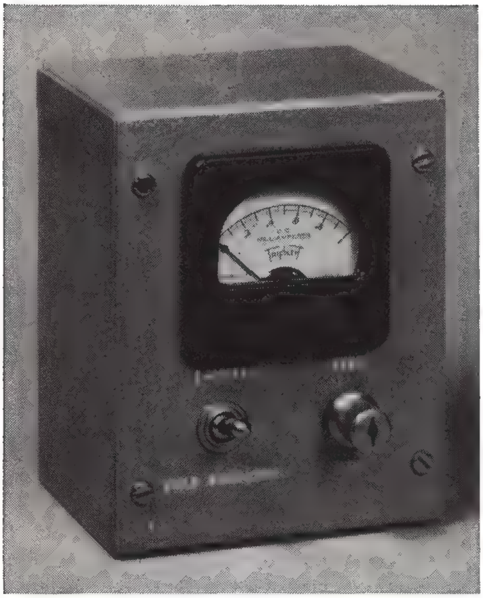


Fig. 15-59—The combination signal-field-strength meter is housed in a 3 × 4 × 5-inch aluminum box. The meter is a 2½-inch rectangular model. R_2 is at the lower right.

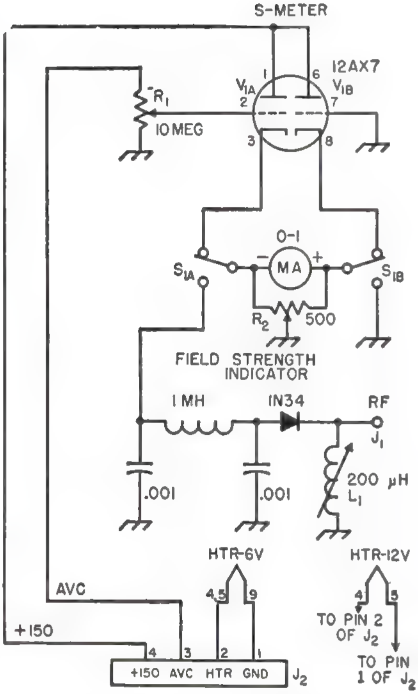
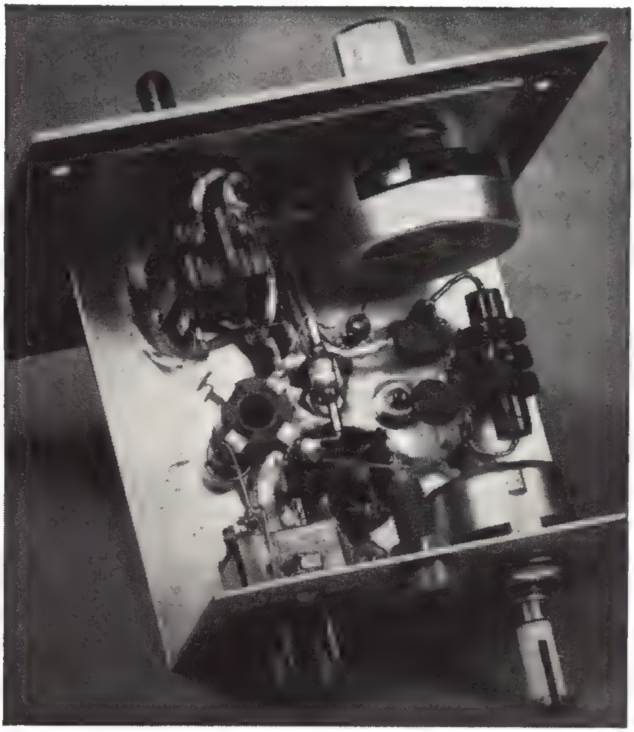


Fig. 15-60—Circuit diagram of the combination field-strength and S meter.

Fig. 15-61—The chassis for the combination indicator is 2⅞ inches wide, 3 inches deep, and 1 11/16 inches high. A ½-inch space is left at the bottom of the panel. S_1 and R_2 are mounted with their centers spaced 1½ inches apart and ⅞ inch up from the bottom of the front edge of the chassis (1⅜ inches from the bottom of the panel). J_2 (Jones P-304AB) and R_1 are similarly centered at the rear, with J_1 in between. L_1, to the left of the tube socket, is a North Hills 120-H, 105 to 200-μh. slug-tuned inductor. The r.f. choke and the two 0.001-μf. capacitors are supported on an insulated lug strip. Holes must be cut in the back cover of the box to clear the projections.



The a.v.c. terminal should be connected to the a.v.c. line in the car receiver, as indicated in Fig. 15-56. It is frequently possible to spot the a.v.c. line by tracing back from the control grid of either the r.f. tube or the converter. The grid of each tube is usually returned to the a.v.c. bus through a resistor of ½ to 1 megohm. If a high-resistance voltmeter is connected between the a.v.c. line and chassis, it should show a negative reading that varies as the receiver is tuned across the broadcast band.

In using the S meter, R_1 should be set at mid-position, and R_2 adjusted to zero the meter with no signal. If strong signals drive the meter off scale, adjust S_1 for less sensitivity, and adjust for zero with no signal again. If the sensitivity is not great enough, increase R_1 and reset to zero with R_2.

When using the field-strength meter, a short length of wire should be connected to J_1. The sensitivity can be varied by R_2 or by adjusting the length of the pick-up wire.

L_1 requires adjustment only in checking on 75 meters. For greatest sensitivity on this band, L_1 should be adjusted for maximum meter deflection. On the higher frequencies the signal strength should be great enough to give adequate meter deflection without the need for a tuned input circuit. Comparative measurements should, of course, be made with the same pick-up wire in the same position relative to the mobile antenna, and with the same adjustment of R_2. The unit, installed in one car, can be used to check the adjustment of other mobile transmitters within a few feet of the pickup.

Magnetic Field Strength Indicator

The field-strength meter of Figs. 15-62 through 15-64 operates on the magnetic component of the electromagnetic field, rather than the electric field. The use of the magnetic field offers certain advantages. The unit may be made more compact for the same sensitivity, and it may be used near an antenna element or other conductor to measure the current flowing in it without cutting the conductor.

The coils are wound on a ferrite rod mounted in a 3 × 4 × 5-inch Minibox with an appropriate tuning capacitor, diode and meter to provide a magnetic pickup shielded from electric fields. The ferrite rod is ¼ inch in diameter and

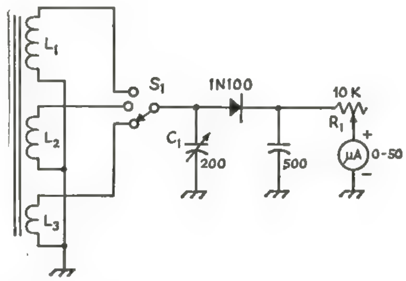


Fig. 15-62—Circuit diagram of the magnetic field-strength meter. Capacitances are in pf.; resistance is in ohms.

C_1—Midget variable capacitor (Hammarlund MC-200).
L_1—3.4 to 12 Mc.—17 turns.
L_2—8 to 25 Mc.—5 turns.
L_3—19 to 40 Mc.—3 turns.
Note: Above coils are close-wound with No. (22 enam.) wire on a single ferrite from ¼-inch diameter, 4 inches long, with no spacing between windings. (Form is antenna rod Cat. No. CF-501, Indiana General Corp. Keasbey, N. J. In case of difficulty in securing this item, one may be obtained from W2OCM* for $1.00 handling expenses.)
R_1—Linear control.
S_1—Ceramic rotary switch: 1 section, 1 pole, 3 positions (Centralab PA-2000).

* 13 Shelley Place, Huntington, L. I., N. Y.

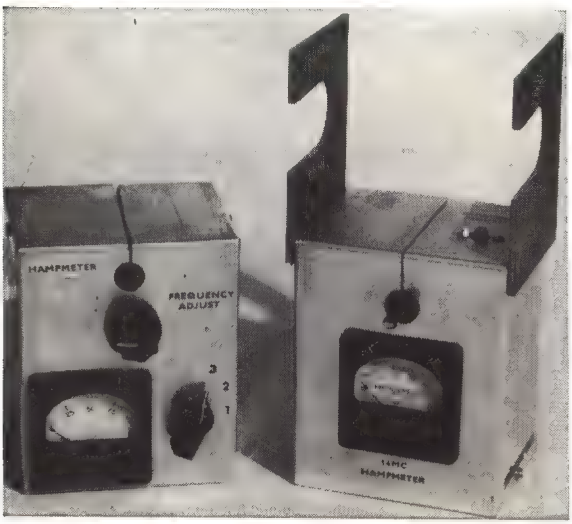


Fig. 15-63—Two versions of the magnetic field-strength meter (K2CU and W2OCM). On the left is the original model covering 3.4 to 40 Mc. in three bands; at the right is a single-band version with hooks for hanging on an antenna.

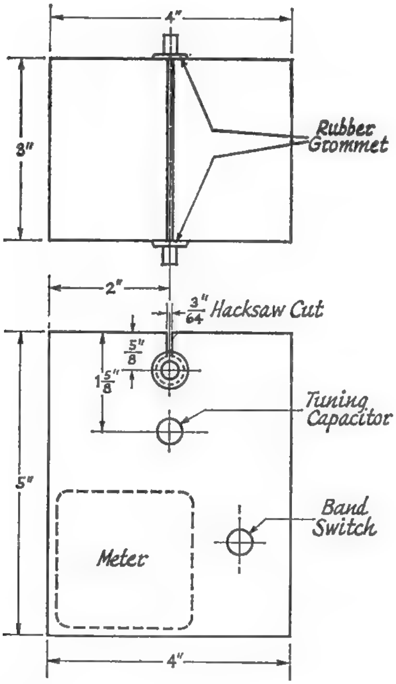


Fig. 15-64—Sketch showing mounting of ferrite through box and saw cut.

4 inches long. It is important to use ferrite of reasonably high μ and Q. Ferramic Q-2 can be used on all amateur bands up through 50 Mc. The next lower grade, Q-1, does not usually work well above about 10 Mc. The winding data accompanying Fig. 15-62 applies to a core of the material and dimensions specified only.

Almost any microammeter can be used, but it is usually convenient to use a rather sensitive meter and provide a large series resistor to adjust the sensitivity as required. The tuning capacitor can be any type that will cover the desired frequency range. If it is provided with a calibrated dial, the instrument can be used as an indicating absorption wavemeter.

The ferrite rod is mounted through the Minibox from front to rear in grommets set in holes about $\frac{5}{8}$ inch from the top of the box. After matching holes have been drilled in the two halves of the box, the box is assembled and a hacksaw cut is made across the top and running down the front and back to reach the holes. This slot prevents the box from acting as a shorted turn.

Hooks made of Lucite, or other similar insulating material, are provided so that the meter can be hung from an antenna element and slid along it to measure the current distribution. In measuring current in a conductor, the ferrite rod should be kept at right angles to the conductor, and at a constant distance from it.

When using the meter, remember that the magnetic field is at right angles to the electric field. In making measurements on a horizontal antenna, the meter should be located at least two wavelengths away from the antenna, at approximately the same height as the antenna, with the ferrite rod in a vertical position. For vertical polarization, the rod should be in a horizontal position, perpendicular to the direction of the antenna.

The meter may be used to detect current flowing in guy wires, masts and towers, coax cable braids, gutters and leaders, and telephone and power wiring, which may have an influence on antenna patterns or be of significance in the case of TVI. These currents may often be eliminated by bonding, or changing the physical lengths involved. When the ferrite rod is oriented parallel to electric-field lines, there will be a sharp null in the reading which may be used to determine the plane of polarization quite accurately.

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